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FIELD DAY

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- IC-12AT Handheld
- NEW! IC-1200 Mobile
- IC-RP1210 Repeater

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The IC-1271A 1240–1300MHz base station transceiver features 10 watts of RF output power, 32 memories, scanning and multi-mode operation including ATV (amateur TV).

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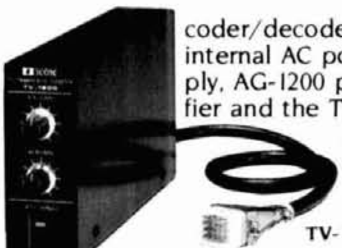

TV-1200

The IC-12AT handheld covers from 1260–1299.990MHz, has ten memory channels, memory scan, program scan and programmable offset. It also features an LCD readout, RIT and VXO, 32 built-in tones and a DTMF pad.

The new IC-1200 1.2GHz mobile transceiver has 21 memory channels, scanning, an HM-14 up/down scanning mic, RIT, large LCD readout and 10 watts power output. Accessories include the PS-45 slim-line external power supply.

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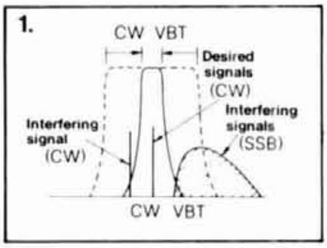
#1 Rated HF!

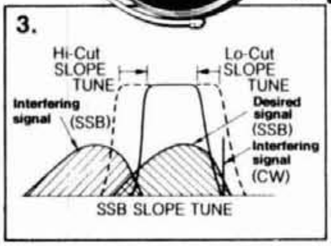
TS-940S

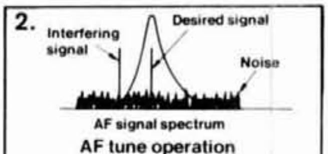
Competition class HF transceiver

TS-940S—the standard of performance by which all other transceivers are judged. Pushing the state-of-the-art in HF transceiver design and construction, no one has been able to match the TS-940S in performance, value and reliability. The product reviews glow with superlatives, and the field-proven performance shows that the TS-940S is "The Number One Rated HF Transceiver!"

- 100% duty cycle transmitter. Kenwood specifies transmit duty cycle ***time.*** The TS-940S is guaranteed to operate at full power output for periods ***exceeding one hour.*** (14.250 MHz, CW, 110 watts.) Perfect for RTTY, SSTV, and other long-duration modes.
- First with a full one-year limited warranty.
- Extremely stable phase lock-ed loop (PLL) VFO. Reference frequency accuracy is measured in ***parts per million!***

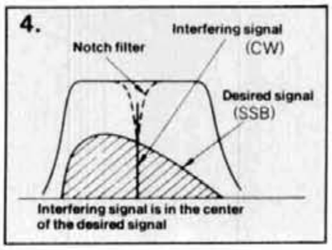






AF tune operation

- The AF TUNE function reduces interfering signals and white noise
- This function should only be used during operation in the CW mode



1) CW Variable Bandwidth Tuning. Vary the passband width continuously in the CW, FSK, and AM modes, without affecting the center frequency. This effectively minimizes QRM from nearby SSB and CW signals.

2) AF Tune. Enabled with the push of a button, this CW interference fighter inserts a tunable, three pole active filter between the SSB/CW demodulator and the audio amplifier. During CW QSOs, this control can be used to reduce interfering signals and noise, and peaks audio frequency response for optimum CW performance.

3) SSB Slope Tuning. Operating in the LSB and USB modes, this front panel control allows independent, continuously variable adjustment of the high or low frequency slopes of the IF passband. The LCD sub display illustrates the filtering position.

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- Superb, human engineered front panel layout for the DX-minded or contesting ham. Large fluorescent tube main display with dimmer; direct keyboard input of frequency; flywheel type main tuning knob with optical encoder mechanism all combine to make the TS-940S a joy to operate.
- One-touch frequency check (T-F SET) during split operations.
- Unique LCD sub display indicates VFO, graphic indication of VBT and SSB Slope tuning, and time.
- Simple one step mode changing with CW announcement.
- Other vital operating functions. Selectable semi or full break-in CW (QSK), RIT/XIT, all mode squelch, RF attenuator, filter select switch, selectable AGC, CW variable pitch control, speech processor, and RF power output control, programmable band scan or 40 channel memory scan.

Optional accessories:
• AT-940 full range (160-10m) automatic antenna tuner • SP-940 external speaker with audio filtering • YG-455C-1 (500 Hz), YG-455CN-1 (250 Hz), YK-88C-1 (500 Hz) CW filters; YK-88A-1 (6 kHz) AM filter • VS-1 voice synthesizer • SO-1 temperature compensated crystal oscillator • MC-43S UP/DOWN hand mic. • MC-60A, MC-80, MC-85 deluxe base station mics. • PC-1A phone patch • TL-922A linear amplifier • SM-220 station monitor • BS-8 pan display • SW-200A and SW-2000 SWR and power meters • IF-232C/IF-10B computer interface.

Complete service manuals are available for all Kenwood transceivers and most accessories.
Specifications, features, and prices are subject to change without notice or obligation.

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contents

on the cover: John Webster, K1FWE (bottom left), and Doug Grant, K1DG (bottom right), operating during the 1986 K1AR Field Day effort. Both are SFRCC members. Top photo: Marty Durham, NB1H, "fixing things" on N1AU's tower.

REFLECTIONS

Novice enhancement and the future of Amateur Radio

Since going into effect last spring, Novice enhancement hasn't caused any great upsurge in the Amateur ranks. Whether it should be considered a modest success or a complete failure seems to depend on who's talking.

If the criterion is merely the decrease in Novice licensees compared with a year ago, Novice enhancement has failed. To me that's a very shallow, superficial interpretation resulting from a cursory reading of the numbers and a misplaced belief that enhancement addressed a basic problem of Amateur Radio rather than an ancillary one.

Discounting the big Novice jump in April and May of 1987, when a lot of newcomers rushed in to take the Novice exam before it was expanded to cover new privileges, the Novice population hasn't shown a significant change in the past year. There are, perhaps, many reasons for that. How many of last year's new licensees didn't even pause at Novice but moved up immediately? How many of them had been putting off becoming Amateurs and were stimulated to take the Novice exam before it got tougher? These and many other questions should be answered before the results of Novice enhancement can be properly assessed.

Some critics now say the problem with Novice enhancement is that it didn't go far enough, and what's really needed is to do away with the CW "boogie man". Though I agree that the CW requirement has long intimidated — and will continue to intimidate — a vocal minority of prospective Amateurs, I also firmly believe that any attack on the CW issue, no matter what its outcome, will have no more effect on the long or short term Amateur growth problem than Novice enhancement did!

Whatever your feelings, neither Novice enhancement nor a no-code license addresses the basic problem. The problem isn't our product, but its marketing. Amateur Radio is a great product, but if our potential customers don't know the product exists, where to find it, or appreciate its many benefits, they aren't going to buy!

Intelligent marketing is based on market analysis. Manufacturers who don't understand this are doomed to slow growth and/or stagnation at best, and the bankruptcy court at worst. Analyses of recent licensees by the FCC and the VEs who are actually bringing the newcomers on board agree that the average new Amateur is an older, well-established adult. Our marketing effort has been aimed at youngsters, so it seems likely we've been targeting the wrong market. The ARRL seems to feel this way, and is now experimenting with a pilot program that encourages older residents of the Tampa/St. Petersburg, Florida area to become hams.

Before investing any great amount of money and effort in new sales pitches or product revisions, I suggest we put some of that money into a professional market study. This study should be directed primarily at those Amateurs who've joined us in the past 10 to 12 years and (when possible) those who've dropped out. It should include questions on how and why respondents became Amateurs, what they felt had helped, or what had hindered their developing Amateur Radio interest. When the study's results are analyzed, the most cost-effective marketing strategy may become clear.

The ARRL and the Amateur Radio Industry Group have the capabilities for such a study. The two worked together well on the *Archie's Ham Radio Adventure* comic book project, and might be willing to work together on this one. In the meantime, however, I feel that any further tinkering with the product isn't going to solve the basic problems, only complicate them.

Joe Schroeder, W9JUV

This editorial is one person's opinion about Novice enhancement and does not necessarily represent the views of *ham radio*. Ed.

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HF transceiver with general coverage receiver.

Compact, easy-to-use, full of operating enhancements, and feature packed. These words describe the new TS-140S HF transceiver. Setting the pace once again, Kenwood introduces new innovations in the world of "look-alike" transceivers!

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- **M. CH/VFO CH sub-dial.** 10 kHz step tuning for quick QSY at VFO mode, and UP/DOWN memory channel for easy operation.
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- **31 memory channels.** Store frequency, mode and CW wide/narrow selection. Split frequencies may be stored in 10 channels for repeater operation.
- **RF power output control.**
- **AMTOR/PACKET compatible!**
- **Built-in VOX circuit.**
- **MC-43S UP/DOWN mic. included.**

Optional Accessories:

- **AT-130** compact antenna tuner • **AT-250** automatic antenna tuner • **HS-5/HS-6/HS-7** headphones • **IF-232C/IF-10C** computer interface
- **MA-5/VP-1** HF mobile antenna (5 bands)
- **MB-430** mobile bracket • **MC-43S** extra UP/DOWN hand mic. • **MC-55** (8-pin) goose neck mobile mic. • **MC-60A/MC-80/MC-85** desk mics.
- **PG-2S** extra DC cable • **PS-430** power supply
- **SP-40/SP-50B** mobile speakers • **SP-430** external speaker • **SW-100A/SW-200A/SW-2000** SWR/power meters • **TL-922A** 2 kW PEP linear amplifier (not for CW QSK) • **TU-8** CTCSS tone unit
- **YG-455C-1** 500 Hz deluxe CW filter, **YK-455C-1** New 500 Hz CW filter.

TS-680S

All-mode multi-bander

- 6m (50-54 MHz) 10 W output plus all HF Amateur bands (100 W output).
- Extended 6m receiver frequency range 45 MHz to 60 MHz. Specs. guaranteed from 50 to 54 MHz.
- Same functions of the TS-140S except optional VOX (VOX-4 required for VOX operation).
- Preamplifier for 6 and 10 meter band.

Complete service manuals are available for all Kenwood transceivers and most accessories. Specifications, features, and prices are subject to change without notice or obligation.

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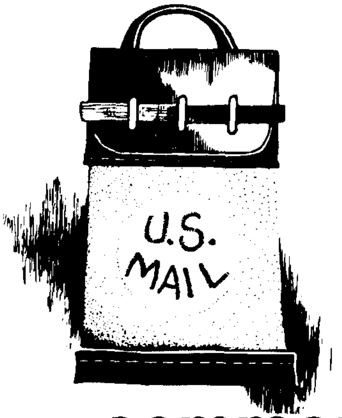

comments

tips for construction projects

Dear HR:

Boy, am I glad that I started building projects *before* reading Paul A. Johnson's article in your March 1988 issue. I'm sure Mr. Johnson's piece would have scared me away. Here are some suggestions for any of your readers who might be interested in project construction:

1. Don't start with anything rf. Receivers, transmitters, tuners, linears are all difficult and require a lot of adjustment once they're assembled. As I recently discovered, even a simple dummy load isn't simple. Don't start with a high-voltage or high-current power supply either. Anything over about 50 volts or 5 amps requires extra care and construction technique. Start with something like a 12 volt 3 amps power supply to run your HT in the house. How about a digital clock? Use a National Semiconductor MA1023C module and matching transformer from Digi-Key and it'll be easy. Get a Curtis chip and make a keyer. These suggested projects may not sound very exciting technically, but you'll find that project construction is often more mechanical than electrical.

2. You don't have to build most projects in metal boxes. For non-rf projects, plastic is fine. It's inexpensive, easy to work, and doesn't have to be painted. Jameco, Digi-Key, and your local Radio Shack all offer a selection of plastic enclosures. Stick with plastic and you won't need a drill press; an electric hand drill is fine. You won't need expensive and dangerous hacksaws, sabre saws, circular saws, or fly cutters either. Holes larger than your drill can make or odd-shaped openings can be cut quickly and easily with a reamer or some cheap files. Filing out openings in metal is an arduous task, but in plastic even cheap files cut quickly. Here's a tip: to drill a nice hole in plastic, start with your smallest bit and work up to the final size using every bit between them in your index. Hot-melt glue guns work on plastic. A cheap pop riveter is another handy tool. If you do need metal, look for a prefab cabinet that will fit your project. Prefabs may seem expensive, but they're a lot easier and you won't need a lot of tools and equipment. Bud, LMB, and others offer an excellent assortment of cabinets ready to house most projects.

3. *Plan! Document!* Much of the work for my projects is done on paper. Start with a good schematic. If you're using any integrated circuits, mark the pin numbers on your schematic. Draw pin diagrams of other parts like transistors next to the part on the schematic. Assign part numbers. Sketch how the project will be assembled, the layout of parts on circuit boards, and the chassis wiring. Then make a from-to wiring list. With all of this planning, your project will be a snap to build and will work the first time. If it doesn't, all the documentation will help you find the problem fast. By the way, keep all of this paper so that if your project ever breaks, you ever want to modify it, or you or a friend ever want to build another, it'll be easy. As you correct bugs or add modifications, document the changes.

4. Take your time. Measure twice, cut once. Make test fits as you move along. Check each electrical connection with an ohmmeter. Try to make every solder joint perfect. Use cable ties or lacing tape to form cable bundles. If you have extensive chassis wiring, use wire marker labels. Use heat-shrink tubing and cable clamps as necessary. In short, try to make each project a show piece inside and out.

While Mr. Johnson's work certainly looks very nice, project construction does not have to be as difficult or as complicated as he makes it sound. You don't have to be a machinist, and you really don't need a lot of expensive tools either. By avoiding complicated projects (especially rf ones) at first, using plastic boxes whenever possible, planning carefully, and working slowly, anyone can enjoy building perfect construction projects. I know I sure do!

Chuck Gollnick, KA7QEN
Ames, Iowa 50010-1363

no contest

Dear HR:

In his article in the November 1987 issue, Bill Orr, W6SAI, says he can attest to the fact that allocation of even a small segment of the 30-meter band to SSB operation would be of great benefit to Amateur Radio. Instead of a bland statement, would he be a bit more explicit?

Surely he must be aware that another well-known author sparked a similar controversy in the columns of the RSGB's *RadComm* magazine, and the consensus of opinion was against any change in the IARU's recommendations. Could there be a conspiracy of authors on this subject?

I have used the 30-meter band almost since it's inception, and the greatest problem is finding a space to work without causing QRM to priority users. Like many others, I have worked over 100 countries; the DX is there and occupancy will surely improve as we advance into the new cycle.

Would the SSB fraternity be as mindful of our non-priority status as the CWers have been? I doubt it, and it would not be long before we lost the band altogether.

I wholeheartedly agree with his 18 MHz sentiments and it would be a great shot in the arm to have the W/Ks on the band, but please *no contests*. As my friend SM3CIQ/U1F says; rather RTTY QRM than contests.

Edward D. Ross, 5B4OG/A9XCE
Larnaca, Cyprus

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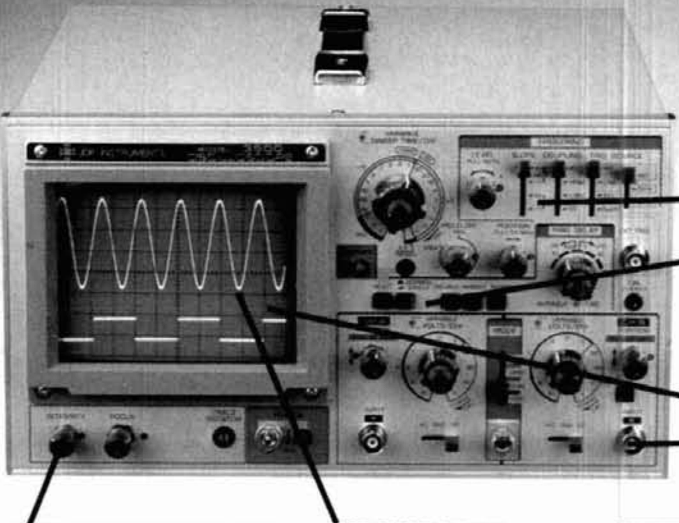

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- Basic DC accuracy: plus/minus 0.25%
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- Transistor Tester: 0°–2000°F
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- Fully overload protected
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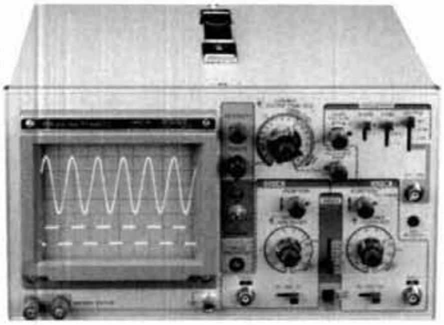

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- Basic DC accuracy: plus/minus 0.5%
- DC voltage: 2v–1000v, 4 ranges
- AC voltage: 200v–750v, 2 ranges
- Resistance: 2k ohms–2M ohms, 4 ranges
- DC current: 2mA–2A, 4 ranges
- Input impedance: 10M ohm
- Fully overload protected
- Approx. 5" x 3" x 1". Under 7 ozs.

DMM-200 $49.95

3.5 DIGIT FULL FUNCTION DMM

Get highly accurate performance at a very affordable price. Rugged construction, 20 amp current capability and 22 ranges make it a perfect choice for serious field or bench work. Low battery indicator and tilt-stand. Probes and 2000 hour battery included.

- Basic DC accuracy: plus or minus 0.25%
- DC voltage: 200mv–1000V, 5 ranges
- AC voltage: 200mv–750V, 5 ranges
- Resistance: 200 ohms–20M ohms, 6 ranges
- AC/DC current: 200µA–20A, 6 ranges
- Input impedance: 10M ohm
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3.5 DIGIT PROBE TYPE DMM

Custom 80 pin LSI chip provides accuracy and reliability in such a compact size. Autoranging, audible continuity and data hold feature help you pinpoint the problem quickly. Case and batteries included.

- Basic DC accuracy: plus/minus 1%
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- AC voltage: 2v–500v, autoranging
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design program
for the grounded-grid 3-500Z

This program gives "no-compromise" answers

There are probably as many amplifiers in existence that use the 3-500Z as there are with any other power amplifier tube. The 3-500Z is an excellent tube with a well-deserved reputation for high power-handling capabilities and reasonable cost. There are no doubt hams with one or two spare 3-500Zs in their shacks who are thinking of building their own amplifiers. But of course, it's one thing to copy the design of another amplifier and a different matter to design one's own from scratch.

Circuit variations for amplifier designs are available from other sources[1] and I will not discuss them here. This article covers only one mode of operation for the 3-500Z — grounded-grid class AB_2 operation — probably the most prevalent use of the tube. I will discuss virtually all possible combinations of drive power, load resistance, drive impedance, plate voltage, and bias requirements for grounded-grid, linear operation. I have included a program which allows you to accommodate *any* set of normal operating conditions that can be realized on the constant-current curves. **Figure 1** is the program listing.

This program's answers are probably a bit more precise than ones obtained with the "Tube Performance Computor".[2] You might assume that linear divisions exist between constant-current curves on the tube charts; they don't. This is not a serious problem, as variations from one tube to another will usually be greater than those differences. It is important to remember that using the hand-calulated methods consumes much paper, time, and nervous energy. My program lets you change any of the input parameters and see the differences for *each* proposed operating condition in a few seconds, as compared with fifteen minutes or a half-hour required for hand-calculated answers — and it doesn't make mistakes!

The program has only 170 lines to enter; it runs completely within a few seconds after you enter the last input. (The answers appear in about 3 seconds on the Tandy-2000, and in 8 to 10 seconds on an early IBM PC.) When this program is compiled, answers appear in under a second. The program is very densely packed with numbers and equations — I know of no other way to define every current and voltage (including fractional values) that can be found on the tube chart, and still use fewer than 7 kilobytes of computer RAM. The labor of typing the program pays for itself many times over as it saves hours of effort during a design routine.

how to use the program

Figure 2 is a reproduction of the "EIMAC 3-500Z Typical Constant Current Characteristics" curves for grounded-grid, class AB_2 operation. The operational area of the program (crosshatching) is superimposed on the curves. Stay inside the "box"; it includes every permissible or useful operating point. You decide on the placement of the operating (or load) line. It is drawn between two points, labeled "i_p", and "Q". The first defines the maximum peak instantaneous plate current as well as the minimum plate voltage. The second defines the quiescent (no drive) value of the plate current, and occurs at exactly the plate-supply voltage. The chart also tells you the quiescent plate dissipation. This is not printed in the program output, but can be calculated by multiplying the resting plate current by the plate supply voltage. The program requests: "Enter Plate Supply Voltage, E_{bb}",

By W.J. Byron, W7DHD, 240 Canyon Drive, P.O. Box 2789, Sedona, Arizona 86336

fig. 1. The 3-500Z Design Program Listing.

```
10 ' SAVED AS 500WRK8
168 PRINT"                     3-500Z Grounded-Grid Characteristics"
170 PRINT"                  Subroutines Copyrighted 1987, W J Byron"
172 PRINT"                                 All rights reserved"
173 PRINT:PRINT
190 INPUT"Enter Plate Supply Voltage, Ebb";E3
191 IF E3>4500 THEN PRINT"        EXCESSIVE PLATE VOLTAGE!":GOTO 190
200 INPUT"Enter Peak Plate Current, Ip";I1
205 IF I1>1.6 THEN PRINT"         EXCESSIVE PEAK CURRENT!":GOTO 200
210 INPUT"Enter Minimum Plate Voltage, Emin";E4
220 IF E4 < 250 THEN PRINT"       HIGH GRID CURRENT AREA!": GOTO 210
221 IF I1>=1.6 AND E4<=1499 OR E4>3000 THEN PRINT"OUT OF BOUNDS!":GOTO 210
222 IF I1>1.4 AND E4<=1499 OR E4>3000 OR I1>1.4 AND E4<1500  THEN PRINT"OUT OF B
OUNDS!":GOTO 210
230 INPUT"Enter Cathode Bias Voltage (Zener)";E2
240 IF E2<0 THEN PRINT"      Negative Cathode Bias Not Permitted!": GOTO 230
290 CLS
300 PRINT"                                   3-500Z"
310 PRINT"                     RADIO FREQUENCY LINEAR AMPLIFIER"
320 PRINT"                         Cathode Driven, Class AB2"
330 PRINT"_____________________________________________________________________"
331 PRINT"Ebb =";E3;"      Ip =";I1;"       Emin =";E4;"       Bias (Zener) =";E2
340 PRINT"_____________________________________________________________________"
350 IF E4=>2000 AND I1>.4 AND I1=<1! THEN GOTO 6000
360 IF E4=>2000 AND I1=<.4 THEN GOTO 5000
370 IF E4=>1500 AND E4=<3000 AND I1>1! AND I1=<1.6 THEN GOTO 7000
380 IF E4=>250 AND E4<2000 AND I1>.4 AND I1=<1 THEN GOTO 3000
390 IF E4=>250 AND I1>1! THEN GOTO 4000
400 IF E4=>250 AND E4<2000 AND I1<.4 THEN GOTO 2000
410 '
500 GOSUB 1000
510 I6 = (I(1)/2+I(2)+I(3)+I(4)+I(5)+I(6)+I(7))/12:
520 I7 = (I(1)+1.93*I(2)+1.73*I(3)+1.41*I(4)+I(5)+.52*I(6))/12:
540 GOSUB 11000
600 '
620 N=1
630 FOR AD = 90 TO 180 STEP 15
640 AR = AD*3.14159/180
650 E5 = (-E1+E2)*SIN(AR)
660 E5 = -E5+E2
670 GOSUB 8000
671 IF E4 >=250 AND E4<1500 THEN GOSUB 9000
672 IF E4>=1500 THEN GOSUB 10000
692 N=N+1
693 NEXT AD
694 E1 = -E1
695 I5 = (I4(1)/2+I4(2)+I4(3)+I4(4)+I4(5)+I4(6)+I4(7))/12
697 I8 = (I4(1)+1.93*I4(2)+1.73*I4(3)+1.41*I4(4)+I4(5)+.52*I4(6))/12
698 I9=I7+I8:E7=E3-E4
710 PD = E1*I9/2:PFT=I7*E1/2:P1=(E1-E2)*I8/2:ZI=E1/I9:PI=E3*I6:PO=E7*I7/2:
720 PP=PI+PFT-PO:ZI=E1/I9:ZO=E7/I7:
722 PRINT"Plate Supply Voltage              =                    ";E3;"Volts"
724 PRINT"Cathode Bias (Zener)              =                    ";INT(10*E2)/1
0;"Volts"
726 PRINT"Zero Signal Plate Current         =                    ";INT(1000*I2)
;"mA dc"
728 PRINT"Single-Tone Plate Current         =                    ";INT(1000*I6)
;"mA dc"
730 PRINT"Single-Tone Grid Current          =                    ";INT(1000*I5)
;"mA dc"
732 PRINT"Grid Power Dissipation            =                    ";INT(P1+.5);"
Watts"
734 PRINT"Peak RF Cathode Voltage           =                    ";INT(10*E1)/1
0;"Volts"
736 PRINT"Feed-through Power                =                    ";INT(PFT+.5);
"Watts"
738 PRINT"Grid Drive Power                  =                    ";INT(PD);"Wat
ts"
739 PRINT"Total Cathode Drive Power         =                    ";INT(PD+PFT+P
1);"Watts"
740 PRINT"Cathode Drive Impedance           =                    ";INT(10*ZI)/1
0;"Ohms"
742 PRINT"Power Input                       =                    ";INT(PI);"Wat
ts"
744 PRINT"PEP Power Output                  =                    ";INT(PO);"Wat
ts"
746 PRINT"Plate Dissipation                 =                    ";INT(PP);"Wat
ts"
748 PRINT"Plate Load Impedance              =                    ";INT(ZO);"Ohm
s"
750 PRINT"_____________________________________________________________________"
760 INPUT"Do you want to change an input - Y or N";A$
762 IF A$ = "Y" OR A$ = "y" THEN 764 ELSE END
764 INPUT"Which: Ebb (1), Ip (2), Emin (3) or Bias (4) - Enter Number";B
766 ON B GOTO 768, 770, 772, 774
768 INPUT"New Plate Supply Voltage";E3:GOTO 290
770 INPUT"New Peak Current, Ip";I1:GOTO 290
772 INPUT"New Min. Plate Voltage";E4:GOTO 290
774 INPUT"New Bias Voltage";E2:GOTO 290
1000 '
1041 N=1
1050 FOR AD = 90 TO 180 STEP 15
```

```
1070 AR = AD*3.14159/180
1080 E5 = (-E1+E2)*SIN(AR)
1090 E5 = -E5+E2
1091 GOSUB 8000
1092 IF E6 =>2000 THEN GOSUB 13000 ELSE GOSUB 12000
1100 N=N+1
1110 NEXT AD
1130 RETURN
2000 '
2030 A = 1.17048 + 8.70277E-03*E4 - 8.7317E-07*E4^2
2040 B = -372.2766 + .3693*E4 - 1.9797E-04*E4^2
2050 C = 1192.038 -2.3118*E4 + .0013494*E4^2
2060 D = -1466.98 + 3.5197*E4 - .002135*E4^2
2070 E1 = (A +B*I1 +C*I1^2 +D*I1^3)
2090 GOTO 410
3000 '
3030 A = 34.095 + 1.526857E-02*E4 - 8.4952E-06*E4^2
3040 B = -339.196 + .026538*E4 + 1.6988E-05*E4^2
3050 C = 381.55 - .106071*E4 - 4.7619E-07*E4^2
3060 D = -183.7 + .088096*E4 - 1.26987E-05*E4^2
3070 E1 = (A +B*I1 +C*I1^2 +D*I1^3) + 1.5
3090 GOTO 410
4000 '
4030 A = 15.3333 - .204*E4 + 6.6667E-05*E4^2
4040 B = -171.805 + .5971*E4 - .0001903*E4^2
4050 C = 69.3 - .53895*E4 + .0001759*E4^2
4060 D = -17.3612 + .16667*E4 - 5.5556E-05*E4^2
4070 E1 = (A+B*I1 +C*I1^2 +D*I1^3)
4100 GOTO 410 '
5000 '
5030 A = 2.5971 + 6.17953E-03*E4 + 6.9333E-08*E4^2
5040 B = -197.3858 - .012861*E4 - 2.7028E-06*E4^2
5050 C = 297.248 + .0315373*E4 + 2.440133E-05*E4^2
5060 D = -231.3046 - .0085463*E4 - 4.3113E-05*E4^2
5080 E1 = A +B*I1 +C*I1^2 +D*I1^3
5100 GOTO 410
6000 '
6030 A = 59.3 - .015733*E4 + 1.467E-06*E4^2
6040 B = -400.519 + .09939*E4 - 6.04467E-06*E4^2
6050 C = 430.4 - .142*E4 + .0000084*E4^2
6060 D = -179.1328 + .0650273*E4 - 3.85547E-06*E4^2
6080 E1 = A +B*I1 +C*I1^2 +D*I1^3
6100 GOTO 410
7000 '
7030 A = -1102 + 1.0133*E4 - 1.9867E-04*E4^2
7040 B = 1855.833 - 1.89027*E4 + 3.7722E-04*E4^2
7050 C = -1106.25 + 1.16875*E4 - .0002375*E4^2
7060 D = 197.9143 - .23264*E4 + 4.86107E-05*E4^2
7070 E1 = (A +B*I1 +C*I1^2 +D*I1^3)
7090 GOTO 410
8000 '
8060 M = (E3-E4)/(E2-E1)
8070 B = E4-M*E1
8080 E6 = M*E5 + B
8100 RETURN
9000 '
9030 A = .028344 - 1.40085E-04*E6 + 6.60444E-08*E6^2
9040 B = -2.855E-05 - 5.972E-06*E6 + 2.52651E-09*E6^2
9050 C = 4.4072E-05 - 6.0547E-08*E6 + 2.1886E-11*E6^2
9060 I4(N) = A +B*E5 + C*E5^2
9061 IF E5>=-24 THEN I4(N)=(-E5/24)*.05
9062 IF E5=>0 THEN I4(N)=0
9080 RETURN
10000 '
10030 A = -.11518 + 1.121096E-04*E6 -2.515184E-08*E6^2
10040 B = -.010834 + 7.15948E-06*E6 - 1.42553E-09*E6^2
10050 C = -5.208347E-05 + 5.38044E-08*E6 - 1.161174E-11*E6^2
10060 I4(N) = .94*(A +B*E5 + C*E5^2)
10061 IF E5>=-24 THEN I4(N)=(-E5/24)*.05
10062 IF E5=>0 THEN I4(N)=0
10080 RETURN
11000 '
11030 A = .0342397 + 1.505585E-06*E3 + 6.804935E-09*E3^2
11040 B = -.0051697 - 3.0065E-07*E3 - 2.15583E-10*E3^2
11050 C = 1.8582E-04 - 1.95465E-08*E3 + 3.351883E-12*E3^2
11060 I2 = A + B*E2 + C*E2^2
11070 IF I2<0 THEN I2=0
11080 IF E2 > (E3+417.5)/150 THEN I2 = 0
11100 RETURN
12000 '
12030 A = .014419 + 2.0981E-05*E6 - 1.11584E-08*E6^2 + 7.9275E-12*E6^3
12040 B = -.0052947 + 2.2872E-06*E6 - 1.2934E-09*E6^2 - 1.4654E-13*E6^3
12050 C = 7.4159E-05 - 4.2928E-08*E6 + 1.17198E-10*E6^2 - 4.8095E-14*E6^3
12060 D = 4.0815E-07 - 1.1512E-09*E6 + 1.5312E-12*E6^2 - 5.1342E-16*E6^3
12070 I(N) = (A +B*E5 +C*E5^2 +D*E5^3)*.985
12090 RETURN
13000 '
13030 A = -.090832 + 1.0172E-04*E6 - 1.1786E-08*E6^2
13040 B = -9.896401E-03 + 2.5728E-06*E6 - 5.7887E-10*E6^2
13050 C = 2.6724E-05 + 2.5902E-08*E6 - 1.5488E-12*E6^2
13060 D = -1.8581E-07 + 1.30725E-10*E6 + 1.184E-14*E6^2
13070 I(N) = A +B*E5 +C*E5^2 +D*E5^3
13071 IF I(N)<0 THEN I(N)=0
13090 RETURN
```

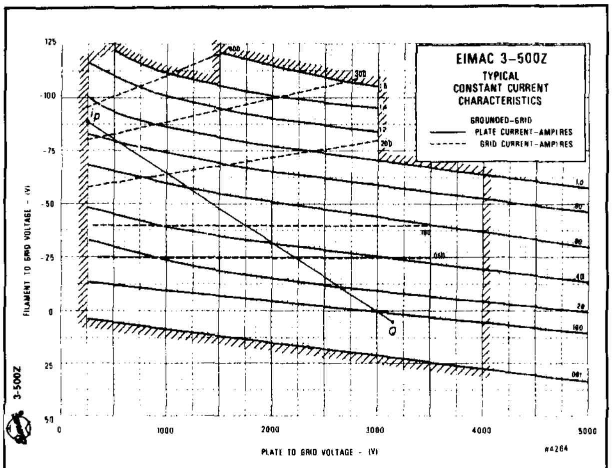

fig. 2. The 3-500Z Constant Current Characteristics Curves with the program boundaries marked. Stay inside the cross-hatched area. The load line is for the solution in *fig. 4*.

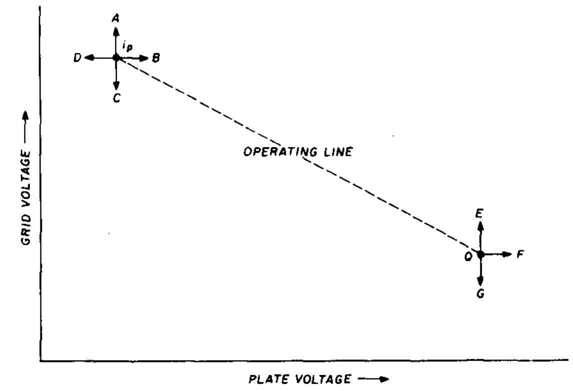

fig. 3. Diagram of the possible movement of the Operating Line. For use with *table 1*.

Table 1. The effects of moving points i_p and Q in the designated directions.

Movement in direction	Results of movement
A:	Higher grid current Higher plate current Higher input and output Higher plate dissipation Lower drive impedance
B:	Reduced efficiency Lower input and output Reduced grid current Increased plate current
C:	Reduced grid current Reduced Plate current Reduced input and output
D:	Higher distortion (peak flattening) Increased grid current
E:	Lower distortion Higher quiescent dissipation
F:	Increased input and output Increased efficiency Increased plate dissipation (Do not exceed mfgr's max Plate Voltage)
G:	Lower quiescent plate current Lower quiescent plate dissipation Increased distortion (non-linear "crossover")

"Enter Peak Plate Current, i_p", "Enter Minimum Plate Voltage, E_{min}", and "Enter Cathode Bias Voltage (Zener)".

These four inputs define both ends of the operating line (which you may already have drawn on the curves), and are sufficient to determine all operating parameters. It isn't actually necessary to draw the line, but it may help to visualize it; one appears in the figure to demonstrate the method.

The main program starts immediately after the last input and calculates a total of fifteen lines of data. Two are input repetitions (Plate Supply Voltage and the Zener Bias); the rest are results of internal calculation by the program. These are the numbers you want. The inputs are repeated in the line just below the heading for the program output as a record of what has been entered. Use them as starting points for any changes you want to make. After the listing there is a question: "Do you wish to change an input—Y or N?" If you enter "Y", the program in turn will ask you, "Which one?" in a menu, and you can change any one of the four inputs until the outputs are to your liking. Any other entry, including "N", will abort the program and you will have to "RUN" again. You can then use the "immediate mode" of BASIC to calculate, for instance, the quiescent plate dissipation (which as a rule of thumb should be somewhere between 30 and 40 percent of the maximum plate dissipation — 500 watts for this tube).

There are some constraints imposed on the initial inputs (see lines 190 through 240). These conditions do not exist at the "Do you wish to change an input?" prompt. Because you are typing the program youself, you decide whether or not to include them (use your own good judgement). To do so, just duplicate the conditions stated in the lines identified above. The constraints result partly from the maximum values permitted by the manufacturer and partly from my work to limit both maximum grid dissipation and amplifier distortion. The program isn't valid outside these limits; the manufacturer's allowable values are the principal reasons for the lower limit of 250 volts for the minimum plate voltage.

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Ebb = 3150	Ip = .87	Emin = 250	Bias (Zener) = 5.2

Plate Supply Voltage	=	3150 Volts
Cathode Bias (Zener)	=	5.2 Volts
Zero Signal Plate Current	=	67 mA dc
Single-Tone Plate Current	=	278 mA dc
Single-Tone Grid Current	=	90 mA dc
Grid Power Dissipation	=	6 Watts
Peak RF Cathode Voltage	=	87.8 Volts
Feed-through Power	=	19 Watts
Grid Drive Power	=	25 Watts
Total Cathode Drive Power	=	50 Watts
Cathode Drive Impedance	=	152.2 Ohms
Power Input	=	878 Watts
PEP Power Output	=	617 Watts
Plate Dissipation	=	279 Watts
Plate Load Impedance	=	6806 Ohms

Do you want to change an input - Y or N?

fig. 4. Program output for the design for two parallel 3-500Zs with 3150 plate supply volts, and 100 watts total drive.

Table 2. Comparison of calculated and manufacturer's data.

Parameter	Program Calc.	Mfgr. Data	Program Calc.	Mfgr. Data
Plate Supply Voltage	1500	1500	3500	3500
Cathode Bias	0	0	+15	+15
Quiescent Plate Current	51	65	25	53
Single-Tone Plate Current	403	400	396	400
Single-Tone Grid Current	142	130	129	108
Grid Drive Power	51	49	49	46
Cathode Impedance	134	94	137	115
Power Input	605	600	1386	1400
PEP Output	330	330	991	890
Plate Load Impedance	1602	1600	5001	5000

points "i_p" and "Q" are moved. **Figure 3** is a schematic showing what to expect when shifting the points. Use it in conjunction with **table 1**. It is essential that you be familiar with these principles — that's the only way you will accomplish your final design.

how well does the program work?

The proof is in the performance. A sample of the program output appears as **fig. 4**, which is also the demonstration of a design using two parallel 3-500Zs. It looks just like the manufacturer's list of typical operating data.

Table 2 compares the program-calculated data with those published by EIMAC under the heading "Typical Operating Data". The results of trying to match two sets (at 1500 volts and 3500 volts) are compared in the table. The manufacturer's data came from the latest technical data sheets for the 3-500Z (the revision effective April 1, 1986).[3] I have culled all except the directly comparable data from the table; they are remarkably close. **Figures 5** and **6** are the program outputs for the 1500- and 3500-volt cases.

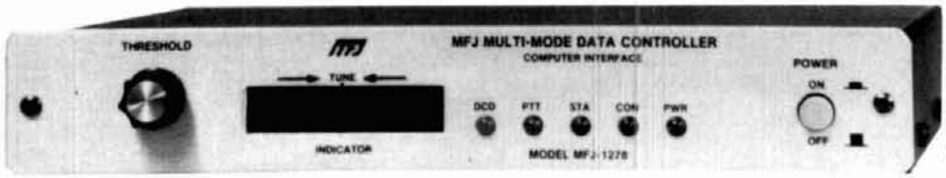
parallel operation

All the tables reflect data for one tube. If you choose a two-tube parallel operation, all currents and power levels must be doubled. All impedances (such as the cathode drive and plate load impedances) must be halved. Voltages remain unchanged.

A hypothetical design demonstration follows:

Suppose you have a power transformer that will deliver 3500 volts dc at no load. A typical power supply voltage will sag about 10 percent under load, so enter 3150 for E_{bb}. Now suppose that you have 100 watts of drive (PEP) from the exciter, and you also want to have the most power available from the amplifier.

MFJ multi-mode data controller



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You can copy all shifts and all standard speeds including 170, 425 and 800 Hz shifts and speeds from 45 to 300 baud. You can *copy not only amateur RTTY but also press, weather and other exciting traffic.*

A high performance modem lets you copy both mark and space for greatly improved copy under adverse conditions. It even tracks slightly drifting signals.

You can transmit both narrow and wide shifts. The wide shift is a standard 850 Hz shift with mark/space tones of 2125/2975 Hz. *This lets you operate MARS and standard VHF FM RTTY.*

You get both the American Western Union and the international CCITT character sets, Autostart for unattended reception and selectable "Diddle".

A receive Normal/Reverse software switch eliminates retuning and Unshift-On-Space reduces errors under poor receiving conditions.

ASCII

You can transmit and receive 7 bit ASCII using the same shifts and speeds as in the RTTY mode and using the same high performance modem. You also get Autostart and selectable "Diddle".

CW

You get a Super Morse Keyboard mode that lets you send perfect CW effortlessly from 5 to 99 WPM, including all prosigns -- it's tailor-made for traffic handlers.

A huge type ahead buffer lets you send smooth CW even if you "hunt and peck".

You can store entire QSOs in the message memories, if you wanted to! You can link and repeat any messages for automatic CQs and beaconing. Memories also work in RTTY and ASCII modes.

***A** tone Modulated CW mode turns your VHF FM rig into a CW transceiver* for a new fun mode. It's perfect for transmitting code practice over VHF FM.

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The CW receive mode lets you copy from 1 to 99 WPM. Even with sloppy fists you'll be surprised at the copy you'll get with its powerful built-in software.

You also get a random code generator that'll help you copy CW faster.

Weather FAX

***You'll** be fascinated as you watch WEFAX signals blossom into full fledged weather maps on your printer.* Other interesting FAX pictures can also be printed -- such as some news photographs from wire services.

Any Epson compatible printer will print a wealth of interesting pictures and maps.

Automatic sync and stop lets you set it and leave it for no hassle printing.

You can save FAX pictures and WEFAX maps to disk if your terminal program lets you save ASCII files to disk.

Pictures and maps can be *printed to screen in real time or from disk* on IBM and compatibles with the MFJ-1284 Starter Pack.

You can transmit FAX pictures right off disk and have fun exchanging and collecting them.

Slow Scan TV

The MFJ-1278 lets you exchange pictures with thousands of SSTVers all-over-the-world.

You'll not only see what your ham buddies look like but *you can send your own pictures* to them, too.

You can print slow scan TV pictures on an Epson compatible printer. If you have an IBM PC or compatible you can print to screen in near real time or from disk with the MFJ-1284 Starter Pack.

You can transmit slow scan pictures right off disk -- there's no need to set up lights and a camera for a casual contact.

You can save slow scan pictures on disk from over-the-air QSOs, audio tapes and other sources if your terminal program lets you save ASCII files.

The MFJ-1278 transmits and receives 8.5, 12, 24, and 36 second black and white format SSTV pictures using two levels.

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***Nothing** beats the quick response of a memory keyer during a heated contest.*

You'll score valuable contest points by completing QSOs so fast you'll leave your competition behind. And you can snag rare DX by slipping in so quickly you'll catch everyone by surprise.

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The program was run with initial inputs of 3150 volts, 1.0 A (chosen as a starting point), 250 volts, and 5.2 volts. The Zener was chosen as 5.2 volts because it is about the same bias voltage as that used in the Heath SB-220. It proved to be a good choice. By reducing i_p incrementally via the menu, the calculated drive power was reduced to 50 watts exactly (for one tube). It occurred when the max plate current (i_p) reached 0.87 A (see **fig. 4**). By "doubling and halving," the resulting numbers for two tubes in parallel are:

Plate Supply Voltage = 3150 volts
Cathode Bias = 5.2 volts
Zero Signal Plate Current = 134 mA dc
Single-Tone Plate Current = 556 mA dc
Single-Tone Grid Current = 180 mA dc
Feed-Through Power = 38 watts
Total Cathode Drive = 100 watts

```
------------------------------------------------------------------------
Ebb = 1500        Ip = 1.31        Emin = 470        Bias (Zener) = 0
------------------------------------------------------------------------
Plate Supply Voltage                   =                      1500 Volts
Cathode Bias (Zener)                   =                      0 Volts
Zero Signal Plate Current              =                      51 mA dc
Single-Tone Plate Current              =                      403 mA dc
Single-Tone Grid Current               =                      142 mA dc
Grid Power Dissipation                 =                      14 Watts
Peak RF Cathode Voltage                =                      117.9 Volts
Feed-through Power                     =                      38 Watts
Grid Drive Power                       =                      51 Watts
Total Cathode Drive Power              =                      103 Watts
Cathode Drive Impedance                =                      134.4 Ohms
Power Input                            =                      605 Watts
PEP Power Output                       =                      330 Watts
Plate Dissipation                      =                      312 Watts
Plate Load Impedance                   =                      1602 Ohms
------------------------------------------------------------------------
Do you want to change an input - Y or N?
```

fig. 5. Program output for duplicating the manufacturer's data for 1500 plate supply volts.

```
------------------------------------------------------------------------
Ebb = 3500        Ip = 1.25        Emin = 350        Bias (Zener) = 15
------------------------------------------------------------------------
Plate Supply Voltage                   =                      3500 Volts
Cathode Bias (Zener)                   =                      15 Volts
Zero Signal Plate Current              =                      25 mA dc
Single-Tone Plate Current              =                      396 mA dc
Single-Tone Grid Current               =                      129 mA dc
Grid Power Dissipation                 =                      11 Watts
Peak RF Cathode Voltage                =                      116.6 Volts
Feed-through Power                     =                      37 Watts
Grid Drive Power                       =                      49 Watts
Total Cathode Drive Power              =                      97 Watts
Cathode Drive Impedance                =                      137.1 Ohms
Power Input                            =                      1386 Watts
PEP Power Output                       =                      991 Watts
Plate Dissipation                      =                      430 Watts
Plate Load Impedance                   =                      5001 Ohms
------------------------------------------------------------------------
Do you want to change an input - Y or N?
```

fig. 6. Program output for duplicating the manufacturer's data for 3500 plate supply volts.

Cathode Drive Impedance = 76 ohms
Power Input = 1756 watts
PEP Power Output = 1234 watts
Plate Load Impedance = 3403 ohms

The actual power output would be (1234 + 38) or 1272 watts because of the feed-through power. Total plate dissipation would be 559 watts. The design can proceed from here.

Always keep the manufacturer's maximum ratings in mind. Two appropriate values to monitor are the plate dissipation and the maximum plate voltage. Another more important one is grid dissipation, also calculated by the program. Normally you will never exceed all the maximum ratings at once — but stay alert to assure that it doesn't happen. This program should be accompanied by the manufacturer's tube data sheets.

comments

In **table 2** there are two lines which show some discrepancies; neither of these is very important. They involve plate quiescent current and cathode drive impedance. These numbers agree with those calculated by hand from the EIMAC Bulletin No. 5 "Tube Performance Computor". Even so, the **table 2** "worst-case" discrepancy (cathode drive impedance for the 1500-volt case) would result in a VSWR of only 1.4. Everything else came out much better than I expected.

My program is complicated and takes lots of time. But remember that when the work is done, the resulting program will simulate data for an operational circuit. It should be emphasized that "real" tubes may produce numbers which differ by as much as ± 10 percent in the main part of the characteristics curves. Though I have made many design calculations with the aid of this program, I certainly have not challenged every possible entry. I would be interested in your results. Let know if there are any "glitches" in the program, and send me your suggestions for improvements.

I plan to put several design programs on disk. I have done this same routine for the 3CX1200A7; the routines for the 8877 are half-finished. I will tackle each tube type in succession until most of the common tubes are covered. In the meantime, I hope you have good results with this 3-500Z design program.

acknowledgment

Thanks to Frank Chess, K3BN, who helped with the programming. It was his idea to echo the inputs at the heading of the output routine.

references

1. W.J. Byron, W7DHD, "Design an Amplifier Around the 3CX1200A7," *ham radio*, December 1987, page 33.
2. Bulletin Number 5, "Tube Performance Computor," Varian/EIMAC, 301 Industrial Way, San Carlos, California 94070
3. Available from Varian/EIMAC, 1678 Pioneer Road, Salt Lake City, Utah 84104

ham radio

designing a station for the microwave bands: part 2

A complete 10-GHz Amateur SSB/CW station

Part 1 discussed why the microwave Amateur bands may be better than lower frequencies for many applications, though in the past Amateurs have viewed them as line-of-sight realms. It described some of the inherent advantages microwaves have for point-to-point communication, even over modern higher power hf, VHF, or UHF stations. These advantages make them very attractive for high volume, high data rate communications like those required for Amateur networking.

A local oscillator frequency scheme using common pc boards was presented. It can be used to get a station on all of the Amateur microwave bands with a minimum of redundant construction. This scheme uses conventional lower frequency components, readily available microwave oscillators, and only a small amount of additional microwave construction to produce a high quality narrowband station. Part 1 and the rest of this series demonstrate this approach by describing construction of a complete 10-GHz Amateur SSB/CW station — the station that holds one end of the current North American 10-GHz DX record of 414 miles.

spectral purity

The cornerstone of this station is a spectrally pure and stable 1010-MHz oscillator. Spectral purity, sometimes overlooked more than it should be even on the hf bands, is of particular importance when operating on microwave frequencies. This is because the "contamination" produced by angular (phase or frequency) modulation of a low-frequency reference signal is multiplied right along with the signal itself when a harmonic is used in a microwave system. The fact that drift and frequency errors are multiplied is well known to anyone who tries to "net" a pair of fm transceivers on 1200 or even 440 MHz. However, these frequency domain "imperfections" are members of a whole class of impurities given the name "phase noise". Even a quartz oscillator in a modern hf transceiver exhibits this to some degree. In a well-designed oscillator the "cleanliness" of a signal is related to its operating frequency. On the Amateur hf bands these noise characteristics may be so small relative to normal signal-to-noise ratios that they are unobservable, except to-noise ratios that they are unobservable, except perhaps as an increase in background noise level down the band from a local "big gun". Some of the early synthesized ham band transceivers exhibited this as noise "humps" a few kHz either side of the carrier frequency on both transmit and receive. Commercial Amateur equipment has improved to the point where fundamental overload or other factors usually come into play before the phase noise of the local oscillators is observed. However, as higher frequencies are required and higher harmonic multiples of reference oscillators are used, these unwanted components are multiplied. The relative amplitude of these unwanted signals follows a 20 log N rule, where N is the harmonic number. This means that on the tenth harmonic of a signal, the phase noise sidebands can be expected to increase by 20 log 10 or 20 dB. The 100th harmonic will be 40 dB worse than the fundamental. Consequently, a "clean" signal at 10 MHz, one with say −90 dBC (dB relative to the carrier) noise sidebands or fm spurious signals, might be 60 dB worse at 10 GHz, or −30 dBC. On an S9 signal such noise might be barely audible; however, if the fundamental oscillator was only −60 dBC the resulting microwave signal might be unusable for communications. Because the 1010-MHz oscillator and its harmonics provide a local oscillator signal for a narrowband station, spectral purity must be maintained. Although a PLL can serve to "clean up" a poorer oscillator at frequencies close in to the carrier, no improvement is made

By Glenn Elmore, N6GN, 3528 Deerpark Drive, Santa Rosa, California 95404

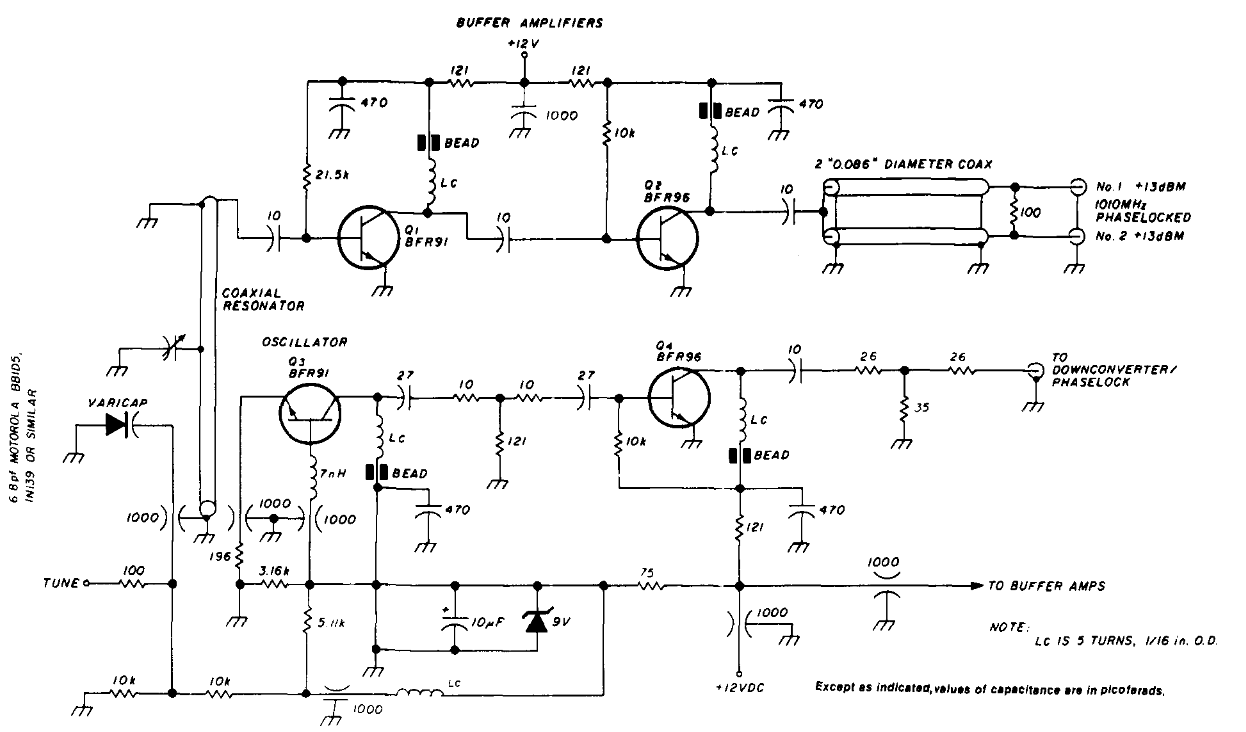


fig. 1. The 1010-MHz oscillator circuit uses a coaxial resonator tuned with a voltage variable capacitor and coupling inductor. The oscillator transistor drives the phaselocking circuitry through a buffer amplifier. Two outputs for signal mixing or microwave PLLs are taken from the other end of the half-wave resonator after two stages of amplification.

beyond the PLL bandwidth. For this reason the best available oscillator should be used.

1-GHz reference oscillator

This oscillator can be used as the LO for a 1296-MHz signal mixer directly, as well as for the microwave harmonic downconverter reference at microwave frequencies. The active device is an inexpensive bipolar transistor. A coaxial resonator is made from pc board and brass tubing. Three separate buffered outputs are provided for phase locking, downconverter reference, and 1296-MHz signal mixer LO. The oscillator is tuned with the same UHF TV tuner diode used in the 100-MHz reference along with a short length of wire coupled to the resonator at the low-impedance end. This provides approximately ± 3-MHz tuning range around a 1010-MHz center frequency.

Oscillator tuning is somewhat novel; it works in much the same way as "loop modulation" of early radio days. Free running high-power oscillators were used with a carbon microphone connected across a single-turn loop located in the vicinity of a frequency determining inductor. As the operator spoke into the microphone the resistive load across the loop varied, which in turn modulated the loop current. Because this was an induced current, it tended to produce an opposing flux which effectively varied the net tank inductance and frequency modulated the oscillator. The technique works, but be careful not to couple too closely or extract so much energy from the tank that you burn up the microphone — not to mention the operator!

The method used here doesn't extract much power from the tank, as the load the varicap presents to the loop is mostly reactive. Any such dissipation is undesirable as it acts to lower the operating Q of the resonator. The varicap value and coupling wire inductance are chosen to be below self-resonance for any tuning voltage. This is done to limit the maximum current and control energy loss in the tuning circuit resistances. If the tuning circuit tunes too close to resonance, oscillation may stop. With nominal loop dimensions and the indicated varicap, the 1010-MHz oscillator tunes with a nearly straight frequency/voltage tuning curve. The 5-MHz tuning range is ample to maintain lock once the other loops and coarse tuning are adjusted to center the output frequency.

Two versions of this oscillator have been built. The first uses a quarter-wave line allowing physically smaller construction, but requiring a dielectric support for the high-impedance end of the line to obtain the lowest "microphonics". The second approach uses a

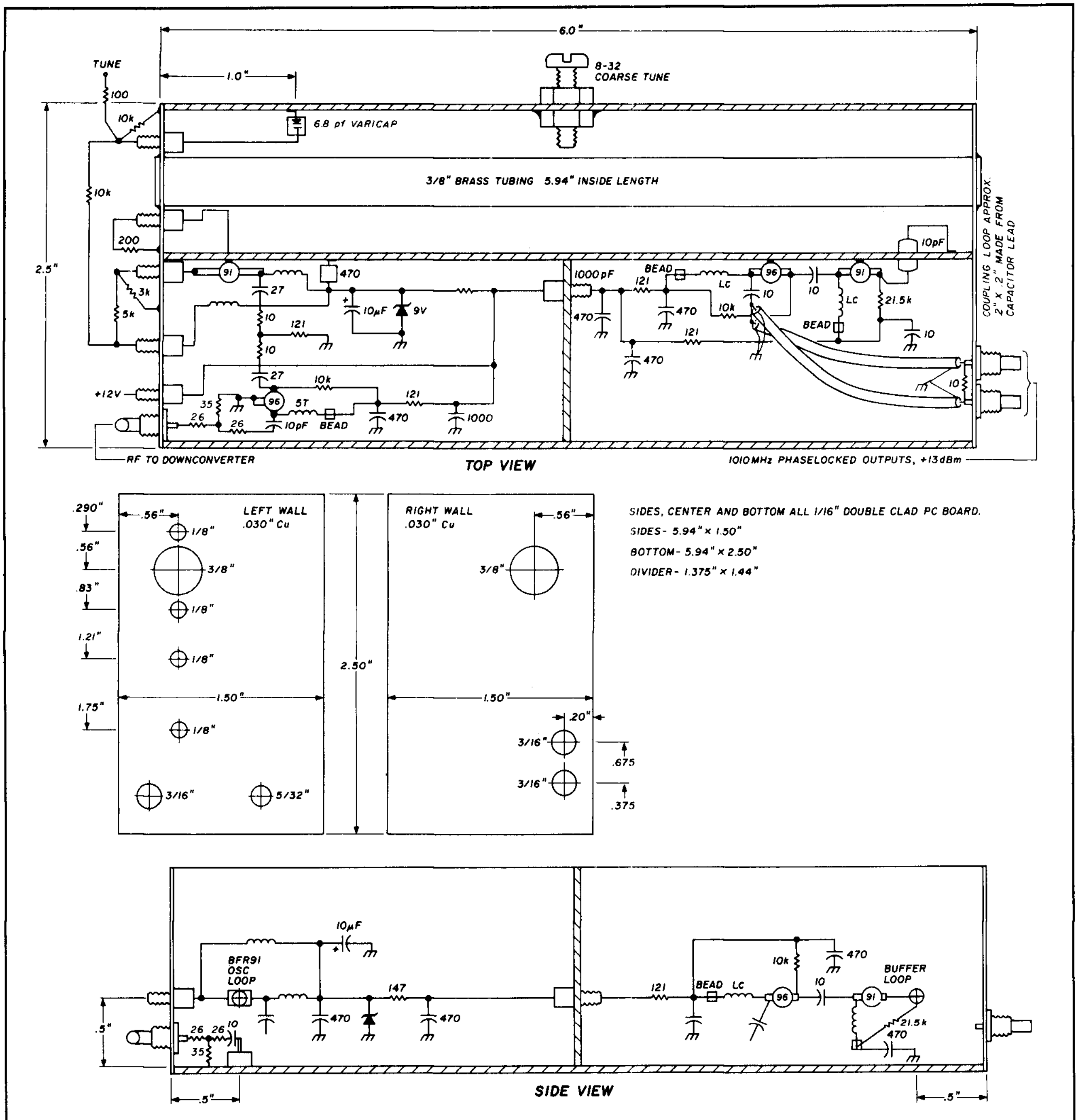


fig. 2. The 1010-MHz oscillator is built from pc board and copper flashing. The assembly is divided into three compartments — one for the resonator and one each for the oscillator and signal amplifier circuits. Minimum lead length is used when components are soldered directly to the circuit board material.

half-wave line and, although longer, is simpler to construct. The quarter-wave version allows tuning versatility by "telescoping" the inner conductor with a length of the next smaller size brass tubing sliding through the center of the fixed tubing. It tuned it continuously from 800 to more than 1200 Mhz.

The half-wave version has another advantage. When you place the oscillator transistor with its isolation amplifier on one end of the resonator and signal amplifiers on the other end, the resonator serves to isolate spurious signals which might be present in the downconverter/phase lock circuitry. This "autofiltering" makes it easier to achieve −80 dBC spectral purity at 1 GHz. Similar performance can be obtained with the quarter-wave version, but more stages of isolation and careful shielding are required.

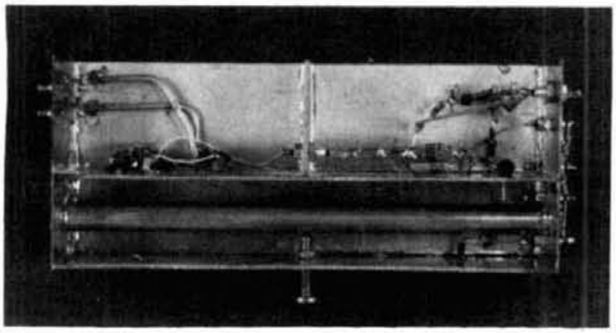


fig. 3. 1-GHz oscillator.

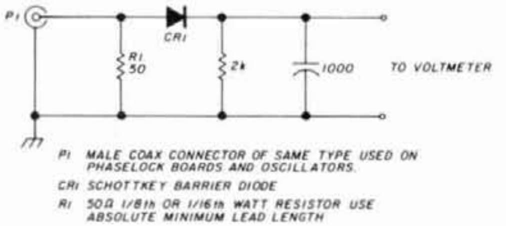


fig. 4. A simple diode detector and voltmeter can measure signal power up through the VHF range. The circuit is useful for relative power measurement well past 1 GHz.

The BFR91 oscillator transistor is optimized to have maximum negative resistance at 1 GHz with the insertion of approximately 7 nH of inductance in its base lead. This inductor is just the 3/8 inch of lead length between the transistor package and the feedthrough capacitor ground on the end wall. The emitter is coupled into the resonator with a loop, also bypassed in a feedthrough capacitor on the same wall. Base and emitter biasing resistors are connected on the outside. The 1-GHz oscillator schematic is shown in **fig. 1**. **Figure 2** shows the mechanical dimensions and positioning for the resonator, feedthrough capacitors, and coupling loops. A photo of the completed oscillator is shown in **fig. 3**.

construction

The oscillator could be built entirely of pc board, but I chose to make the end walls from copper flashing. This makes it easier to solder the brass tubing after the resonator box has been assembled. The sides, center, ends, and partition should all be punched or drilled before soldering. Holes for the oscillator emitter loop and buffer amplifier input loop are in the center wall. Amplifier transistor emitters and all bypass capacitors can be soldered directly to the board material with virtually no excess lead length. The oscillator emitter lead can protrude right through the center wall hole and be soldered to its coupling loop. An 8-32 brass nut should be soldered to the inside wall of the resonator so that a tuning screw can be inserted later. If possible use 1/8th or 1/16th watt resistors. The physically smaller packages should have less associated inductance. Choose feedthrough capacitors small enough to fit snugly against the brass tubing on the oscillator end. These must be soldered in place since their nuts would otherwise interfere with the brass tubing protruding from the end wall.

adjustment

Begin check-out without tuning screws and apply 12 volts. The oscillator emitter (measured at the outside of its feedthrough capacitor) should sit at about 3.5 volts, and the amplifier transistors should have 6 to 10 volts on their collectors. Collector currents of about 15 mA for the BFR91 and 40 mA for the BFR96 amplifiers are fine. All three outputs should have a load connected; a 50-ohm resistor may be tacked across an unused output as a temporary load. If a power meter or other calibrated detector is not available, an inexpensive power detector may be made (**fig. 4**). An approximate calibration curve useful through the VHF range is shown in **fig. 5**. At 1 GHz the curve may not accurately predict the detected power because of

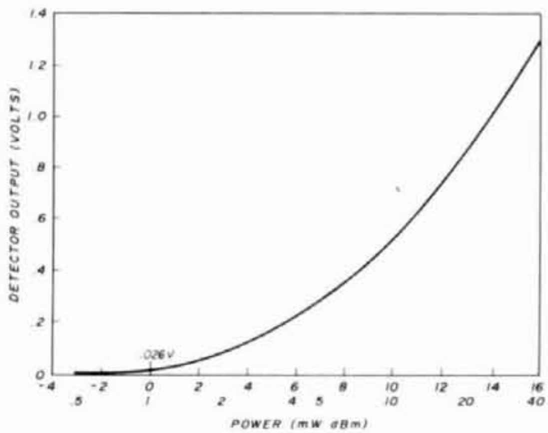


fig. 5. A plot of the detector output voltage as a funtion of input power shows a useful range from about 0 to +16 dBm (1 to 40 milliwatts).

differing construction techniques and component characterisics, but the detector should still be useful for determining relative output powers and adjusting the 1010-MHz circuits. I built the detector right on the cable end of the same type of SMB coax connector I used throughout. You can use it to verify ECL outputs as well as oscillator performance.

A 1-GHz frequency counter or a spectrum analyzer is extremely useful for tune-up. If such test equipment is not available, build the 1-GHz harmonic downconverter described in the next section. Use it to convert the 1-GHz signal down to the hf range of a general coverage receiver or low frequency counter. If you use a receiver, couple the downconverter lightly or use an attenuator to avoid overload. Overloading can cause confusion because of images and other spurious responses.

With an applied fixed tuning voltage of 6 volts, insert the tuning screw and set the frequency to approximately 1010 MHz. Adjust the emitter loop slightly to assure oscillation while varying the tuning voltage over the 2-to-10 volt varicap tuning range. Reduce coupling by decreasing the area of the loop and positioning it further from the brass tubing. Use the minimum coupling to maintain output so you can avoid unnecessarily loading the resonator and degrading phase noise. This coupling is somewhat dependent on resonator loading by both the tuning circuit and buffer amplifier input loops. Adjust the buffer amplifier loop (made from the coupling capacitor lead) for minimum coupling consistent with maximum power out of the power splitter. Adjust the emitter loop to maintain output over the whole tuning range. Some iteration between these two adjustments may be necessary to arrive at the best settings. If you find that the oscillator dies at the high end of the tuning range, or just above 10 volts, you may need to lower the tuning circuit resonant frequency. Do this by lengthening the tuning inductor slightly. The values shown in the drawing should provide a good starting place and should work without modification. Extreme emitter loop overcoupling can cause "squegging", the output switching rapidly between two frequencies. This is not a problem if the above adjustment procedure is followed. Reduce coupling if you observe spurious sidebands on the unlocked oscillator or find low-frequency oscillations on the bias feedthrough capacitors.

The output amplifier on the signal side is followed by a power splitter made from two 2-inch lengths of semi-rigid coax. This is a simple way to provide two outputs. If only one 1010-MHz source is required, it may be omitted and the single BFR91 buffer amplifier used to provide +10 dBm for a signal mixer. The two-stage amplifier with a BFR96 in the output and the power divider can easily provide two +13 dBm (20 milliwatt) sources.

Once the loops are positioned for proper power output, all that remains is to readjust the tuning screw so the oscillator "free runs" right at the desired frequency. If you adhere to the dimensions for the half-wave version, the oscillator should run at about 1025 MHz with no tuning screw and only the 4.5 volts from the resistive divider on the tuning input. It should tune down mechanically to 1000 MHz without a significant change in output power. When you obtain the proper frequency, secure the tuning screw locking nut. Verify that approximately + and − 2 MHz tuning is produced with 10 volts and 2 volts on the tuning input, respectively.

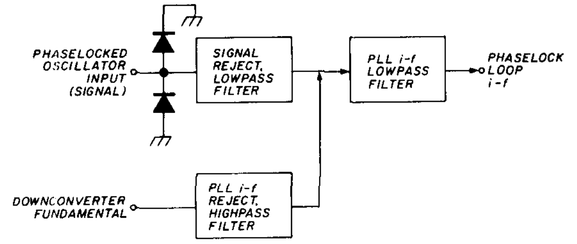


fig. 6. This harmonic downconverter produces an i-f output which is the difference between N multiplied by the downconverter fundamental and the phaselocked oscillator input, where N is an even number. With inexpensive diodes and Amateur construction techniques, i-f output power of −30 dBm is readily available.

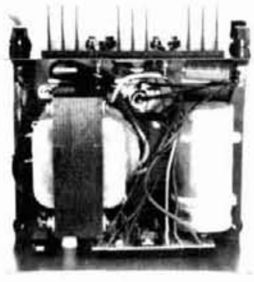
PLL harmonic downconverters

The downconverters themselves are similar, although implementation at 10 GHz is somewhat different from that at 1 GHz. Anti-parallel diodes are used with a diplexer arrangement to couple signals in and out. The downconverter block diagram is shown in **fig. 6**.

The anti-parallel diode pair is effectively an *even* harmonic mixer. Its simplicity and built-in protection from overload and static damage make it attractive for this application. Depending on harmonic number and phase-locked oscillator frequency, −30 to −40 dB conversion efficiencies are obtainable even with "ham shack" construction — i.e., discrete components or microstrip circuits cut out of Teflon™ epoxy pc board material with a small hobby knife. The high-pass filter couples the reference fundamental into the diodes; the low-pass filter couples the i-f out. The oscillator can be connected directly to the diode pair through a small capacitor.

At 1 GHz, packaged diodes and discrete capacitors and inductors can be used. Lead length should be kept to a minimum, but otherwise the circuit is extremely simple to build. The diodes generate considerable energy at odd harmonics of 100 MHz. However, the isolation of the 1010-MHz oscillator resonator, not to mention the buffer amplifiers, keeps this energy from showing up in the signal output. These sidebands are for the most part amplitude, not frequency modulated, and don't get "amplified" when higher harmonics of the 1010-MHz signal are used as a reference signal in the 10-GHz downconverter. The PLL i-f signal

ASTRON CORPORATION
9 Autry
Irvine, CA 92718
(714) 458-7277



INSIDE VIEW — RS-12A

ASTRON POWER SUPPLIES

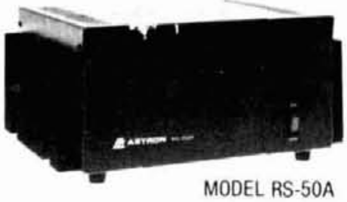
• HEAVY DUTY • HIGH QUALITY • RUGGED • RELIABLE •

SPECIAL FEATURES

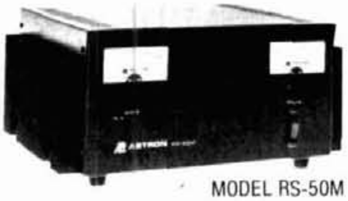
- SOLID STATE ELECTRONICALLY REGULATED
- FOLD-BACK CURRENT LIMITING Protects Power Supply from excessive current & continuous shorted output
- CROWBAR OVER VOLTAGE PROTECTION on all Models **except RS-3A, RS-4A, RS-5A.**
- MAINTAIN REGULATION & LOW RIPPLE at low line input Voltage
- HEAVY DUTY HEAT SINK • CHASSIS MOUNT FUSE
- THREE CONDUCTOR POWER CORD
- ONE YEAR WARRANTY • MADE IN U.S.A.

PERFORMANCE SPECIFICATIONS

- *INPUT VOLTAGE: 105-125 VAC*
- OUTPUT VOLTAGE: 13.8 VDC ± 0.05 volts (Internally Adjustable: 11-15 VDC)
- RIPPLE Less than 5mv peak to peak (full load & low line)
- **Also available with 220 VAC input voltage**



MODEL RS-50A

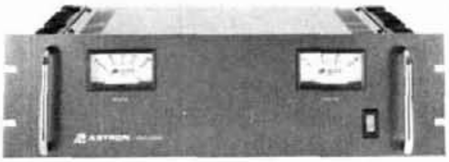


MODEL RS-50M

MODEL VS-50M

RM SERIES

19" × 5¼ RACK MOUNT POWER SUPPLIES



MODEL RM-35M

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RM-12A	9	12	5¼ × 19 × 8¼	16
RM-35A	25	35	5¼ × 19 × 12½	38
RM-50A	37	50	5¼ × 19 × 12½	50
• Separate Volt and Amp Meters				
RM-12M	9	12	5¼ × 19 × 8¼	16
RM-35M	25	35	5¼ × 19 × 12½	38
RM-50M	37	50	5¼ × 19 × 12½	50

RS-A SERIES

MODEL RS-7A

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-3A	2.5	3	3 × 4¾ × 5¾	4
RS-4A	3	4	3¾ × 6½ × 9	5
RS-5A	4	5	3½ × 6⅛ × 7¼	7
RS-7A	5	7	3¾ × 6½ × 9	9
RS-7B	5	7	4 × 7½ × 10¾	10
RS-10A	7.5	10	4 × 7½ × 10¾	11
RS-12A	9	12	4½ × 8 × 9	13
RS-12B	9	12	4 × 7½ × 10¾	13
RS-20A	16	20	5 × 9 × 10½	18
RS-35A	25	35	5 × 11 × 11	27
RS-50A	37	50	6 × 13¾ × 11	46

RS-M SERIES

MODEL RS-35M

MODEL	Continuous Duty (Amps)	ICS* (Amps)	Size (IN) H × W × D	Shipping Wt. (lbs.)
• Switchable volt and Amp meter				
RS-12M	9	12	4½ × 8 × 9	13
• Separate volt and Amp meters				
RS-20M	16	20	5 × 9 × 10½	18
RS-35M	25	35	5 × 11 × 11	27
RS-50M	37	50	6 × 13¾ × 11	46

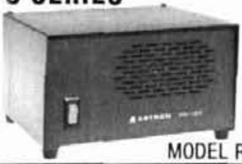
VS-M AND VRM-M SERIES

MODEL VS-35M

- Separate Volt and Amp Meters • Output Voltage adjustable from 2-15 volts • Current limit adjustable from 1.5 amps to Full Load

MODEL	Continuous Duty (Amps) @13.8VDC	@10VDC	@5VDC	ICS* (Amps) @13.8V	Size (IN) H × W × D	Shipping Wt. (lbs.)
VS-12M	9	5	2	12	4½ × 8 × 9	13
VS-20M	16	9	4	20	5 × 9 × 10½	20
VS-35M	25	15	7	35	5 × 11 × 11	29
VS-50M	37	22	10	50	6 × 13¾ × 11	46
• Variable rack mount power supplies						
VRM-35M	25	15	7	35	5¼ × 19 × 12½	38
VRM-50M	37	22	10	50	5¼ × 19 × 12½	50

RS-S SERIES



MODEL RS-12S

- Built in speaker

MODEL	Continuous Duty (Amps)	ICS* Amps	Size (IN) H × W × D	Shipping Wt. (lbs.)
RS-7S	5	7	4 × 7½ × 10¾	10
RS-10S	7.5	10	4 × 7½ × 10¾	12
RS-12S	9	12	4½ × 8 × 9	13
RS-20S	16	20	5 × 9 × 10½	18

*ICS—Intermittent Communication Service (50% Duty Cycle 5min. on 5 min. off)

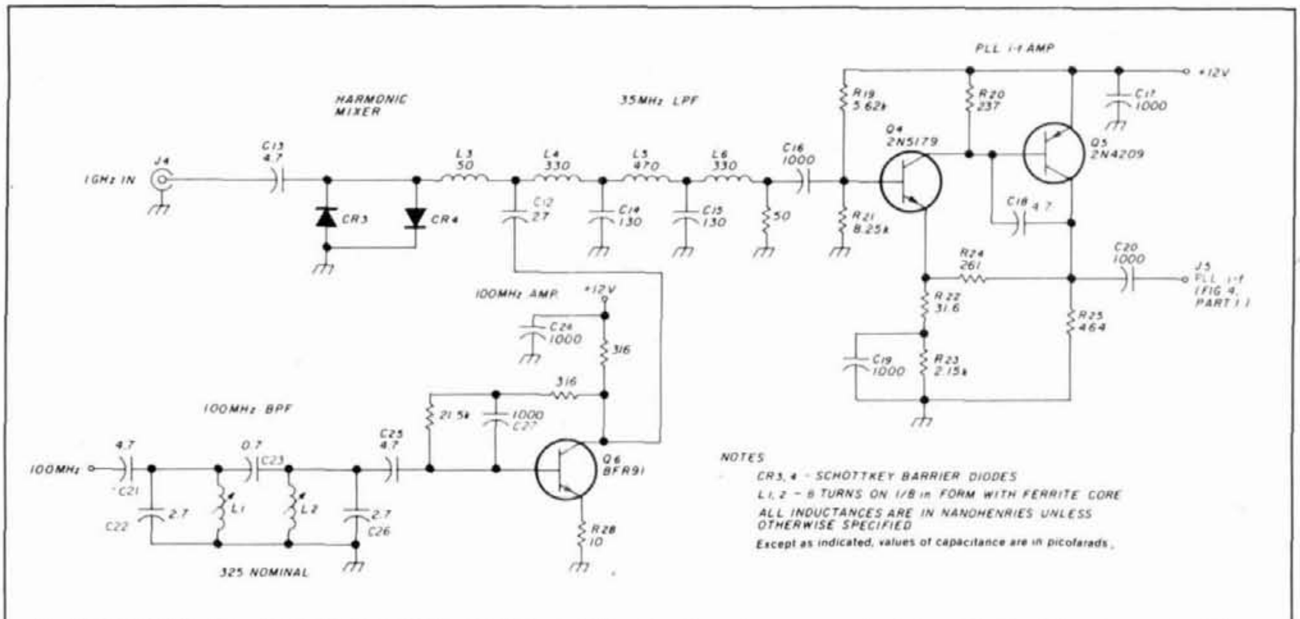


fig. 7. The 1-GHz harmonic downconverter and i-f amplifier circuit produce approximately −10 dBm i-f output when driven with 100-MHz ECL levels and the 1010-MHz oscillator. Leads should be kept short and good VHF practice followed, otherwise no special precautions need to be taken.

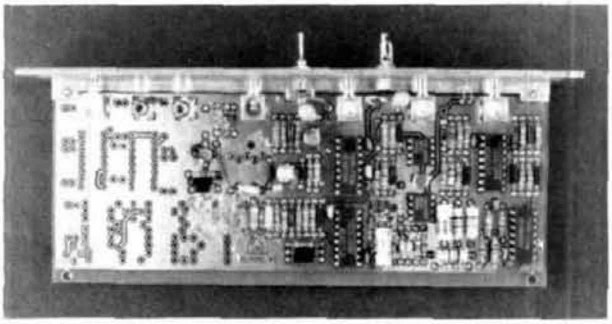


fig. 8. Completed 1-GHz downconverter and common PLL board.

from the downconverter is approximately 30 dB below the reference or locked oscillator levels. This conversion loss is made up for in the bipolar amplifier and the two ECL line receivers on the phase-lock circuit. With 10 to 13 dBm reference drive, i-f output doesn't change dramatcally for 0 to 10 dBm oscillator input. Around −30 dBm PLL i-f power is typical for both converters — plenty to drive the last ECL line receiver before the phase comparator well into saturation. The i-f output may actually drop if the oscillator input level is increased too far. The 1-GHz harmonic downconverter schematic diagram is shown in **fig. 7**. **Figure 8** is a photo of the completed 1-GHz downconverter and common PLL board.

The 100-MHz reference signal is bandpass filtered and amplified from the 0-dBm ECL levels. The filtering makes sure that any low level, low-frequency digital signals which might be present on the 100-MHz ECL output don't "ride" straight through to the PLL i-f amplifiers. Diode drive of 10 to 20 milliwatts is adequate.

The 10-GHz downconverter is functionally the same as the 1-GHz version. Here, however, a pair (or half a quad) of diodes in a small package is used to avoid parasitic inductance and capacitance associated with the larger discrete diodes. Many of the filter elements are made using microstrip techniques instead of lumped components. Chip capacitors are used to minimize parasitic inductance.

Because most of the 10-GHz oscillator power is needed for converting the VHF signal to and from 10 GHz, a hybrid coupler is used to extract only enough to make the PLL downconverter operate. This hybrid has one of its input ports terminated with a discrete resistor. This termination needn't be very good at 10 GHz, as the object of the coupler is simply to extract a sample of the energy (10 dB or so down) and its directivity isn't particularly important. Use as physically small a resistor as possible with 0 lead length. All of the high-impedance lines may be made from some small diameter wire and soldered across the wider traces. Number 38 wire should be fine for this.

The signal mixer is shown with the 10-GHz downconverter and can be built on the same board at the same time. This makes it possible to get on the band as soon as the 10-GHz oscillator is locked and a VHF i-f is available. The 10-GHz harmonic downconverter is shown in **figs. 9A** and **9B**. **Figure 10** shows a 2:1

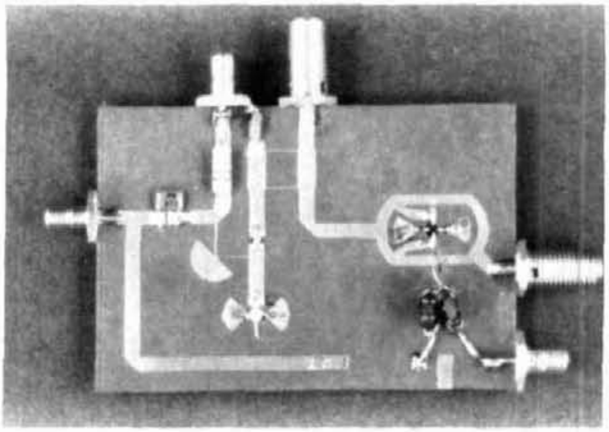
✓ 164

✓ 166

✓ 167

✓ 165

✓ 190



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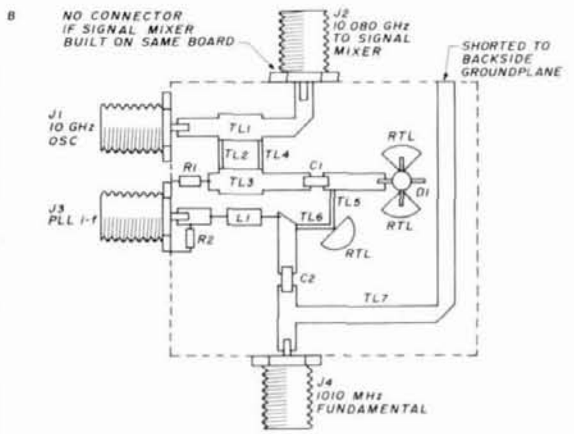
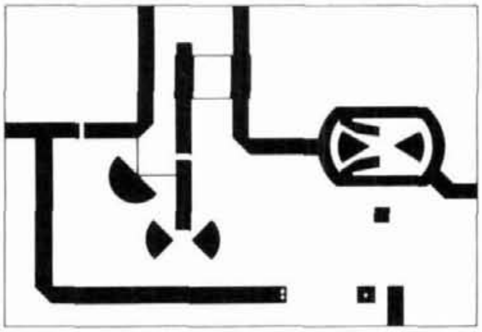


D1 HP HSCH-6812 DIODE QUAD OR METELLICS MSS30-142-E45
TL1, TL3 .100" WIDE x 0.23" LONG
TL2, TL4, TL5, TL6 #36-#40 BARE WIRE x 0.20" LONG
L1 0.1 μH CHOKE
C1 4.7 pF CHIP CAPACITOR
C2 27 pF CHIP CAPACITOR
RTL 0.145" RADIUS QUARTER OR HALF CIRCLE
TL7 AND ALL OTHER LINES 0.083" WIDE
R1, R2 1/16 WATT OR SMALLER 50Ω
J1-J4 MA-COM OR SIMILAR FLANGED SMA CONNECTOR
BOARD MATERIAL 1/32" THICK DOUBLE CLAD DUROID 1.5" x 1.7"
CONNECTOR FLANGES SOLDERED TO BACKSIDE GROUNDPLANE
R1, R2 SOLDERED TO FLANGE OF J3

fig. 9. The 10-Ghz and 1010-MHz downconverters are functionally identical. At 10 GHz, however, microstrip components replace discrete components. A hole is provided in the 1/32 inch Teflon board material under diode ring, D1, to allow shorting the diode leads to the backside ground. Radial transmission lines on these same leads help assure a low-impedance ground connection.



NOTE: BOARD MATERIAL IS 1/32" DOUBLE CLAD TEFLON™/FIBERGLASS. DUROID™ MAY BE USED.

fig. 10. The downconverter/signal mixer assembly is the only microwave circuit that needs to be constructed. The traces may be made using a small hobby knife by the "cut and peel" technique. Tolerances are not extremely critical although a microscope can be a great aid. More detail of the downconverter and signal mixer portions is given in figs. 9 and 19, respectively.

layout of the combined 10-GHz downconverter/signal mixer; **fig. 11** shows a schematic of this board.

locking to 1010 MHz

The 1010-MHz common PLL circuit is nearly identical to that of the 100-MHz reference oscillator — only the loop filter values are different. For this loop, the phase comparator VCO input comes from the filtered and amplified output of the 1-GHz harmonic downconverter. A 35-MHz low-pass filter follows the downconverter; the PLL i-f is first amplified by a two-stage controlled-gain amplifier. I used this configuration instead of another ECL line receiver for two reasons: it allowed variation of the stage gain by changing a single resistor value, and the bipolar amplifier has lower bandwidth than the ECL line receiver. The rest of the PLL circuitry is identical to the 100-MHz phase lock except for the loop filter component values. The bandwidth of this loop is set to approximately 50 kHz.

Once the 1010-MHz oscillator is built and adjusted, you are ready to lock it up. Use one of the common PLL boards with the loop filter component values in part 1, **table 2**. Set the jumper wires on the phase comparator input for the " + " configuration. If the PLL board is working properly (remember that you can test it ahead of time by using it to lock up the 100-MHz oscillator), the loop should close and "pull in" the 1010-MHz oscillator exactly on frequency. This lock can occur if the 100-MHz loop is locked or free running, and the output frequency should be exactly 10.1 times the 100-MHz crystal oscillator frequency. Make sure you use the 10-MHz reference to lock at 1010 MHz and the 20-MHz reference if you are trying to lock to 1020 MHz.

Troubleshoot any problems by checking the PLL board and the 1010-MHz oscillator independently of each other. As long as the oscillator tunes over the correct range and the PLL board is working, there

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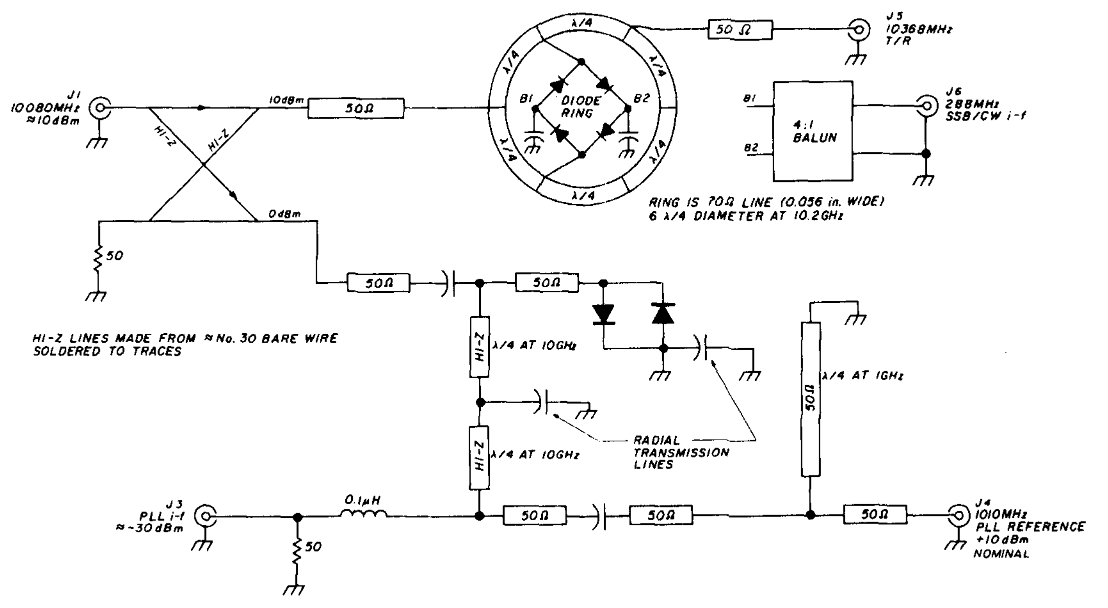


fig. 11. The microwave board mainly uses distributed elements. Impedance of the microstrip transmission lines is controlled by width variation. The high-impedance lines can be made from bare No. 30 wire soldered right to the traces; this is easier than trying to cut or etch 0.005-inch wide traces.

should be no difficulty in achieving and maintaining lock. Once this is done, you have an LO for use in a 1296/2304 station or as a reference oscillator for locking your 10-GHz oscillator.

10-GHz oscillator selection and locking

The 10-GHz oscillator is locked in the same manner as the 1010-MHz reference. The tuning circuit may depend on the type of oscillator available. Generally, only enough tuning range to overcome drift and instability is used. If too much tuning range is provided, the microwave oscillator might get on the "wrong side" of the downconverter reference frequency harmonic, giving an i-f with the wrong tuning sense. If this happens, the PLL amplifier tries to tune the oscillator in the wrong direction to acquire phase lock and the loop will remain saturated and unlocked. For a 20-MHz PLL i-f, 30 MHz of total electronic tuning range should be adequate, and this combined with about a 10-volt swing out of the loop amplifier suggests a 3-MHz/volt tuning sensitivity. If the microwave oscillator is unstable or drifts (necessitating a greater tuning range), an ECL divide-by-2 or divide-by-4 could be inserted right at the phase comparator input. Of course, this would produce a different locked output frequency, and all other i-f and oscillator frequencies in the system might have to be reselected. The loop filter component values would also have to be recomputed.

Selection of the 10-GHz oscillator depends upon

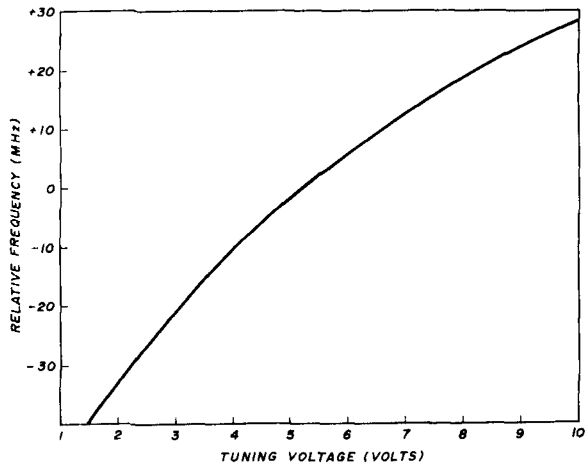


fig. 12. Measured tuning curve for a M/A-Com Gunnplexer™. The curve is fairly straight indicating nearly constant tuning sensitivity, particularly in the 4-10 volt region.

what is available and within your budget. The M/A-Comm Gunnplexers™ work extremely well and require very little additional circuitry. If you have one of these as part of a wideband station, you may want to use the 10,220-MHz locking scheme. If there is already some broadband 10-GHz activity in your area and you

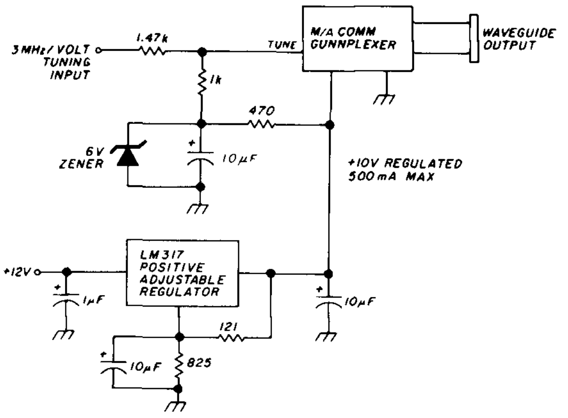


fig. 13. A three-terminal regulator provides clean bias for the Gunnplexer. A Zener diode and two resistors provide an offset and scale the tuning to the approximately 3-MHz/volt sensitivity required for the common phaselock board.

pushing". These are very similar to Gunnplexers except for their lack of electronic tuning and a mixer diode. The tuning deficiency can be overcome by using the bias/tuning circuit in **fig. 15**. Here a three-terminal regulator sets the bias and tunes the oscillator for phase locking. To pick the nominal bias point, plot a frequency versus bias voltage curve for your particular oscillator — this will vary from oscillator to oscillator. Usually a range of bias can be found (often just on one side of maximum power output) that provides a fairly straight tuning curve or nearly constant tuning sensitivity. A plot of a typical bias-tuned oscillator is shown in **fig. 16**. The tuning resistor values are selected to tune over a 24-MHz range with 2 to 10 volts on the tune input. The nominal operating

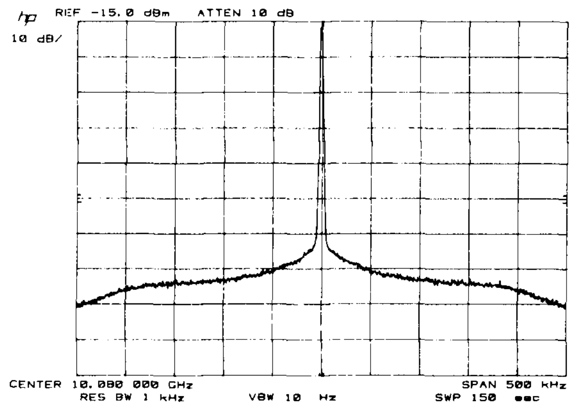
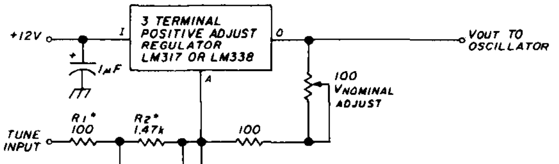


fig. 14. Signal produced by the phaselocked Gunnplexer.

don't want to give it up entirely, this approach will allow switching between modes. The Gunnplexer can be operated with its internal diode mixer for operation on 10220/10250 wideband duplex, or phase locked to 10220 and used with a 148-MHz SSB transceiver for 10368-MHz narrowband weak signal work. The wideband station can also be run phase locked with modulation of the 20-MHz reference signal in the 1020-MHz loop phase. (This should end any local discussions about who is or is not on the right frequency!)

The M/A-Comm Gunnplexers have electronic tuning and need only level shifting and scaling of the tuning voltage. A typical tuning curve for a GunnPlexer is shown in **fig. 12**. Driving the tuning input directly from the loop amplifier provides too much tuning range and could allow "latch-up" on the wrong side of the i-f, as mentioned before. It is a simple matter to scale the tuning input to reduce the approximately 7-MHz/volt sensitivity down to about 3. A circuit providing this scaling, as well as a regulated 10-volt bias supply, is shown in **fig. 13**. This circuit will maintain proper output and tuning even when the power supply voltage drops slightly below 12 volts. A low dropout regulator may be substitued for the LM317K for particularly low inputs. This is of concern primarily when mountain topping with discharged batteries as the only power source! The phase-locked Gunnplexer produces an excellent 10080-MHz signal (**fig. 14**).

Some means of tuning must be provided if an oscillator without an electronic tuning input is used. The Gunn oscillators in automatic door openers can be made to work by using bias voltage "frequency



ΔV24MHz VOLTS	R1 OHMS	R2 OHMS	C1 (NON POLARIZED) µF
0.5	750	10k	0.15
1	330	4.7k	0.33
2	150	2000	0.68
3	82.5	1100	1.5
4	50	680	2.2
5	27	390	4

fig. 15. An adjustable three-terminal positive regulator can both bias and tune a surplus Gunn oscillator via "frequency pushing", the oscillator's frequency/bias voltage dependency. The table shows alternate component values for different "push-tuning" sensitivities. Part of the voltage setting and tuning resistance is bypassed to reduce noise on the regulator output.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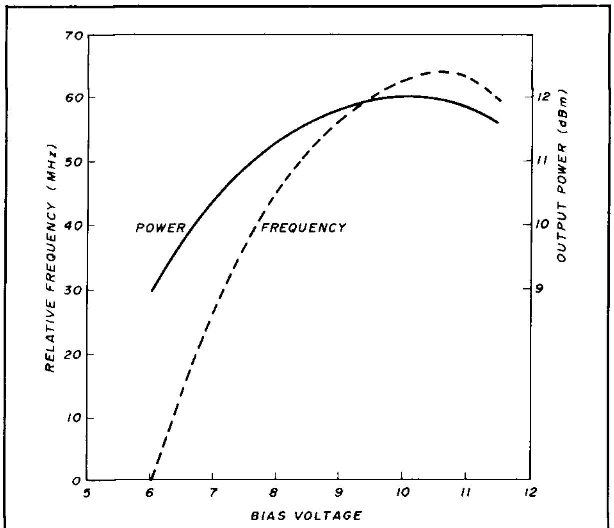

fig. 16. This is a tuning curve of a surplus Solfan™ oscillator of the type used in burglar alarm motion detectors and automatic door openers. Both output power and frequency are dependent upon bias voltage. By plotting a similar curve and selecting a useful portion of the tuning curve, you can find component values for biasing and tuning almost any similar oscillator. In this case, a bias of 8.25 volts + 1.25 volts will tune the output over approximately a + 12-MHz tuning range.

point, with 6 volts applied to the tuning input, is set at the center of this range. If 24-MHz tuning is not possible, use the maximum available and recalulate the PLL component values for the different tuning sensitivity.

The three terminal regulators work in this application because they have several hundred kHz of bandwidth and can follow a 50-kHz bandwidth error signal without adding much additional phase shift. This is necessary for the loop to remain stable. The regulators do add some noise to the oscillator output when used in this configuration; reduction of this is the reason for splitting up and bypassing part of the voltage setting and tuning resistances. This technique is not the ultimate in low phase noise performance, but the −90 dBC noise sidebands obtainable (1 Hz bandwidth) are more than adequate for Amateur use and will probably never be observed unless signal strengths are 30 or more dB above S9. The Gunnplexers, with their built-in tuning, will probably be at least 8 to 10 dB cleaner than this. Although I have not tried them, many of the oscillators in automotive radar detectors should work well. Another source of suitable oscillators is the type used for police radar guns. The NEC ND751AAM for 10 GHz (ND610AAM for 24 GHz) has similar characteristics. Any of these 10-GHz sources should have adequate drive power for the signal mixer described next.

The oscillators' output and antenna connections need to be in coax in order to use the downconverter and mixer. Although waveguide is well behaved and very low in loss, coax is versatile and convenient. I have used coax throughout the 10-GHz station, both at 1 and 10 GHz. Miniature SMB "snap on" connectors work well at 1 GHz and below, even when used with poor quality lossier coax cable. In the microwave region, 0.086-inch semi-rigid cable is a pleasure to work with; the cable and corresponding SMA connectors are fairly easy to find. To cut the cable to length, first score the outer conductor with a sharp knife; then grab each side of the score mark with a pair of needle-nose pliers and break. The Teflon dielectric can be trimmed away and the cable end slid into the connector or soldered directly to the circuit, depending upon the application.

If your oscillator is similar to the door-opener type, it probably has a waveguide output and will require a waveguide-to-coax adaptor. These are often available as surplus but if you don't have or can't get one, it is easy to build an acceptable substitute. The version shown in **fig. 17** made from a short length of commercial waveguide works very well, although you'll need metal-working equipment. If your shack doesn't include much more than a soldering iron, hacksaw, and file, the second version made from pc board in **fig. 18** is for you.

After selecting your 10-GHz oscillator, build the appropriate tuning circuit. If a means of measuring 10-GHz frequency (a 10-GHz counter or spectrum analyzer with 1-MHz frequency resolution) is not avail-

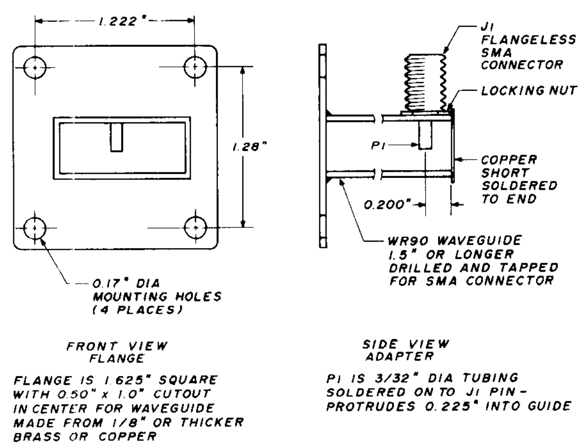

fig. 17. A waveguide/coax adapter which provides very low loss and better than a 1.2:1 VSWR can be made from a short length of standard WR90 waveguide. A standard SMA connector is threaded into a hole in the broad wall of the guide located 0.200 inch from the shorted end. A flange made from a suitably stiff piece of brass or copper is soldered to the other end.

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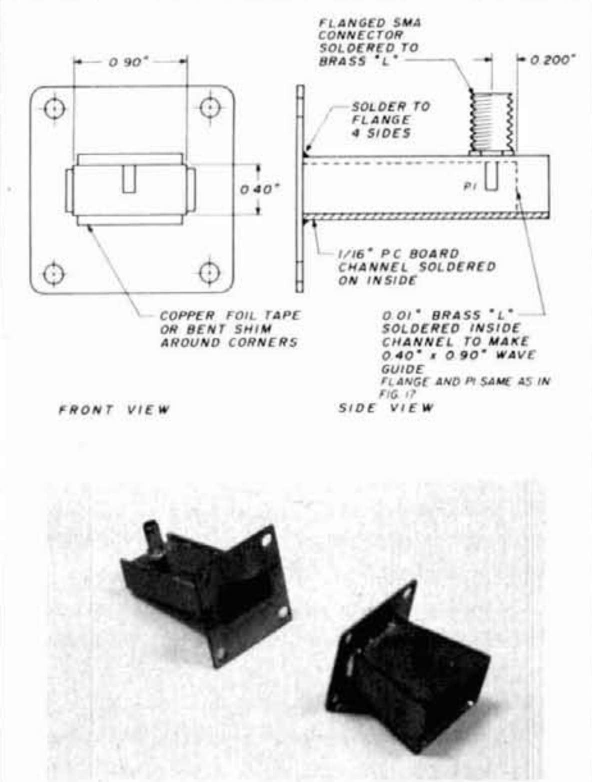


fig. 18. If metal-working equipment necessary to fabricate the adapter shown in *fig. 17* is not available, an acceptable substitute can be made from standard 1/16-inch pc board and brass shim stock. A channel is first made from the board material and then the brass top/short is soldered. Continuity from the inside of this homebrew waveguide to the front of the flange is provided by sticky copper foil tape.

able, you may use the 10-GHz downconverter and locked 1010-MHz source to determine oscillator operating frequency. An old general coverage receiver with poor selectivity is great for this, because an unstable signal is easy to hear as it drifts past. Take the same precautions mentioned earlier to avoid overload. When you do hear a signal, verify that it is on the correct side of the 1010-MHz reference harmonic at 10.1 GHz by making sure that the i-f signal is tuning lower as you tune the microwave oscillator higher. Use mechanical tuning for this because (unless you have a M/A-Comm or other "known" oscillator) you can't be sure what the sense of the electronic tuning is. If the tuning characteristics are unknown, use your general coverage receiver or low-frequency counter on the PLL i-f to plot a tuning curve as a function of the oscillator bias voltage. A few oscillators will "mode" and jump frequency, particularly when not properly matched and operated at extremely low or high biases. A nominal operating point between 6 and 10 volts is appropriate for most oscillators I have tried. Select the nominal bias voltage as the center of a reasonably straight 24-MHz range near the maximum power bias point, or in a mode-free region. Select the tuning scaling resistors from **fig. 15** based on the change in bias necessary to produce 24-MHz frequency change; this gives 3-MHz/volt sensitivity at the tuning input. The sense of this tuning may be either positive or negative, depending on your particular oscillator. For the motion detector oscillator plotted in **fig. 16** I chose a nominal bias point of 8.25 volts. Tuning resistor values were selected for the required volt change (approximately 2.5). These values cause the 2 to 10 volts from the loop amplifier to tune the oscillator over a 24-MHz total range. Resistor and capacitor values for some different oscillator sensitivities are shown in the table.

When you are confident that the oscillator is tuning correctly, preset it to 10080 MHz with 6 volts on the bias circuit tuning input. Do this by coarse tuning for a 20-MHz PLL i-f on the correct side of 10100 MHz. If the PLL board is functioning, locking should now be no more difficult than locking the 1010-MHz oscillator. Remember to select the proper wire jumpers based on "high side" LO and the tuning direction of your particular oscillator.

As for the 1010 MHz oscillator, troubleshoot any problems by separating the PLL components and testing them individually. Make sure that the PLL board works on a lower frequency loop. Verify that there are suitable 1-volt peak-to-peak ECL levels on both phase

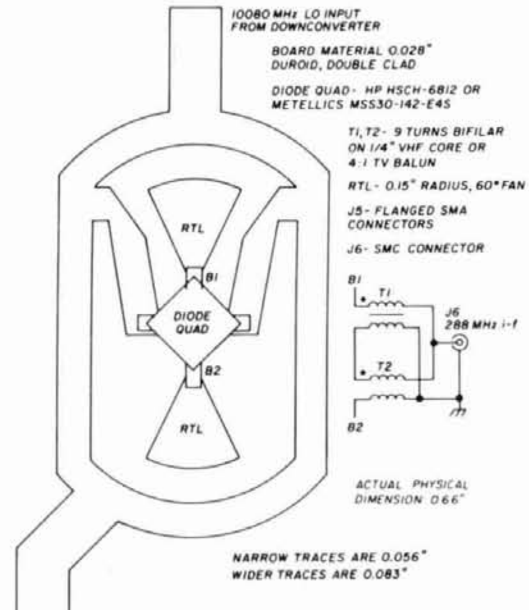


fig. 19. Closeup of the 10-GHz signal mixer.

comparator inputs. Also make sure that the 10-GHz oscillator is tuning properly. Be sure that there is no large (bigger than 1000 pF) bypass capacitor across its bias input; this could limit the fm bandwidth.

signal mixers

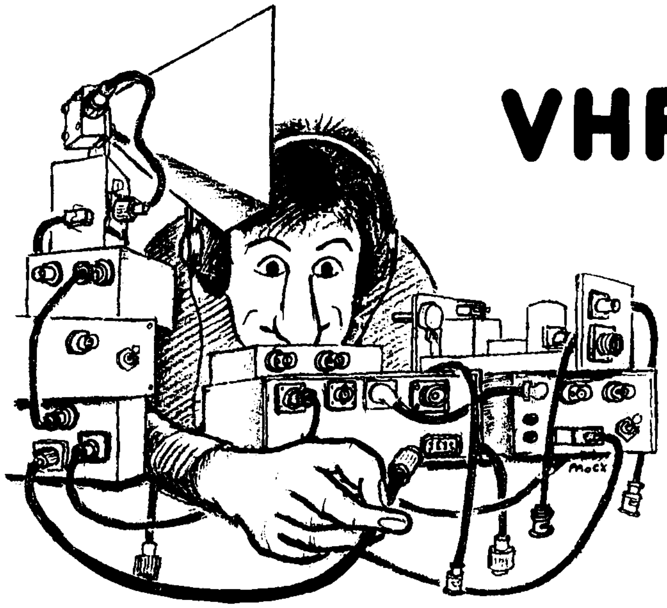
Commercial mixers that give good performance up through 2304 MHz are available at reasonable prices. Simple "rat race" mixers can be made on Teflon pc board for all bands up to and including 10 GHz; they don't work as well at 24 GHz and above because of packaged diode size and parasitics. A diode mixer with less than 7-dB conversion loss at 10368 MHz (with 10080-MHz local oscillator injection) can be cut out of a piece of circuit board. This by itself (no amplifier, preamplifier, or transmit/receive switch) can give S9 signals between similar stations with 4-foot dishes separated by 10 miles!

The 10-GHz signal mixer uses the same diode ring and board material as the 10-GHz harmonic downconverter. Building it on the same piece of board material eliminates two connectors and some coax along with their associated losses. A balun is used to match the mixer diode's i-f impedance to 50 ohms. You can make this balun from two toroidal cores, or use a VHF TV 300-to-75 ohm balun. Conversion loss of under 10 dB should be possible over a range of local oscillator powers. Low barrier diodes are indicated in the parts list, but medium and high barrier may be substituted if sufficient 10080-MHz oscillator power is available. Higher drive levels make higher i-f levels possible on transmit, and therefore higher 10368-MHz transmit power. To avoid serious distortion, i-f power should generally be kept at least 10 dB below the available local oscillator power. A close-up of the 10-GHz signal mixer is shown in **fig. 19**.

Build the signal mixer as part of the downconverter assembly, and you can be on the air as soon as the 10-GHz oscillator is locked and you have a suitable SSB i-f transceiver. Just hook it through a bandpass filter to your antenna!

Part 3 will discuss the following: a 260-MHz locked oscillator along with amplifiers and switching for the 280-290 MHz i-f transverter; and a two-stage, 16-dB gain, 2.5-dB noise figure 10-GHz amplifier that can be used on transmit and receive. Two such stations connected to modest size antennas should improve your DX possibilities and could help you break the current world 10-GHz DX record!

ham radio



VHF/UHF WORLD

Joe Reisert, W1JR

propagation update—part 2

I've discussed VHF/UHF/microwave and millimeter-wave radio propagation many times in this column,[1-6] yet there is always more new material available. This month we will try to pick up where **reference 3** left off and update the present state of the art (SOA) of radio propagation above 50 MHz.

DX records

For years I have felt that the greatest incentives for experimentation on the frequencies above 50 MHz are discovering new propagation modes and setting new DX records. However, published DX records were either scattered or incomplete and often without any mention of the propagation mode used. Most of the published records were worldwide, tending to favor regions where special geography or phenomena are present.

Several years ago I started publishing consolidated VHF/UHF/microwave and millimeter-wave DX records in "VHF/UHF World." At first only the more available worldwide records were included.[1] Later EME (Earth-Moon-Earth) records were added.[2]

I next published a list including only those DX records where at least one of the stations was located in North America. As a new twist, the suspected propagation mode was added.[2] This made for many new DX opportunities above 50 MHz.

The "North America Only" list caught on like wildfire. Many new DX record claims were documented and other propagation modes on different frequency bands were added. These records have been published at least once a year in this column; we now publish new record claims at the end of each "VHF/UHF World."

This month is no exception. All three record tables have been updated. **Table 1** shows the North America Only terrestrial records, **table 2** lists the worldwide terrestrial records, and **table 3** the worldwide EME records.

Each claim has been documented by at least one of the record holders. To facilitate new claims, I designed the VHF/UHF/SHF Record Verification Form in **table 4**. The form verifies when the claimed contact took place and shows the equipment required to make the record. The latter is particularly important since it sets the minimum equipment specifications required.

frequency bands

The list of frequencies available to Amateurs under FCC jursidiction was published in **reference 2**; the microwave and millimeter frequencies were later updated and appeared in **reference 3**. There haven't been any changes of late.

However, there are some further frequency restrictions. The band most affected is the 70 cm (420-450 MHz). Any United States station operating within 100 miles of any PAVE PAWS radar installation and running more than 50 watts is required to obtain FCC permission. This currently affects Amateurs in New England, Georgia, Texas, Alaska, and California.

Amateurs operating in the 70-cm band near the missile test ranges in California, Florida, and New Mexico are also affected by the new rules. There have been additional restrictions placed on Amateurs operating in the 420-430 MHz region near the Canadian border and some Canadians are now affected in the 430-450 MHz region near airports using experimental wind-shear radar. These rules seem to be in a state of flux.

At the present time, Amateur restrictions on 33 cm (902-928 MHz) are

Table 1. North American VHF and Above Claimed DX Records. (Notes 1, 2 & 3)

Frequency	Record Holders	Date	Mode	DX Miles	(km)
50 MHz	Note 4				
EME	WA4NJP (EM84DG)-KH6HI (BL01XH)	88-02-15	CW	4530	(7289)
144 MHz					
Aurora	KA1ZE (FN31TU)-WB0DRL/WA0TKJ (EM18CT)	86-02-08	CW	1347	(2167)
Ducting	KH6GRU (BL01XH)-WA6JRA (DM13BT)	73-07-29	CW	2586	(4161)
EME	VE1UT (FN63XV)-VK5MC (QF02EJ)	84-04-07	CW	10,985	(17676)
Spor. E	KD4WF (EM92LA)-NW70/7 (DM25GV)	87-06-14	SSB	1980	(3186)
FAI	W5HUQ/4 (EM90GC)-W5UN (DM82WA)	83-07-25	CW	1228	(1976)
MS	K5UR (EM35WA)-KP4EKG (FK68VG)	85-12-13	SSB	1960	(3153)
TE	KP4EOR (FK78AJ)-LU5DJZ (GF11LU)	78-02-12	SSB	3933	(6328)
Tropo	K1RJH (FN31XH)-K5WXZ (EM12QW)	68-10-08	CW	1468	(2362)
220 MHz					
Aurora	W3IY/4 (FM19HA)-WB5LUA (EM13QC)	82-07-14	CW	1145	(1842)
Ducting	KH6UK (BL11AQ)-W6NLZ (DM03TS)	59-06-22	CW	2539	(4086)
Spor. E	K5UGM (EM12MS)-W5HUQ/4 (EM90GC)	87-06-14	CW/SSB	932	(1499)
EME	K1WHS (FM43MK)-KH6BFZ (BL11CJ)	83-11-17	CW	5058	(8139)
MS	K1WHS (FM43MK)-K0ALL (EN16NW)	85-08-12	SSB	1279	(2057)
TE	KP4EOR (FK78AJ)-LU7DJZ (GF05RJ)	83-03-09	CW/SSB	3670	(5906)
Tropo	VE3EMS (EN86QJ)-WB5LUA (EM13QC)	82-09-28	SSB	1181	(1901)
432 MHz					
Aurora	W3IP (FM19PD)-WB5LUA (EM13QC)	86-02-08	CW	1182	(1901)
Ducting	KD6R (DM13NI)-KH6IAA/P (BK29GO)	80-07-28	CW	2550	(4103)
EME	K2UYH (FN20QG)-VK6ZT (OF78VB)	83-01-29	CW	11,567	(18612)
MS	W2AZL (FN20VI)-W0LER (EN35IA)	72-08-12	CW	1019	(1640)
Tropo	WB3CZG (FN21AX)-WA5VJB (EM12LQ)	86-11-29	SSB	1318	(2121)
903 MHz					
EME	K5JL (EM15DQ)-WB5LUA (EM13QC)	88-02-07	CW	187	(301)
Tropo	W2PGC (FN02OR)-K3SIW/9 (EN52WA)	86-12-24	SSB	478	(769)
1296 MHz					
Ducting	KH6HME (BK29GO)-WB6NMT (DM12KU)	86-08-13	SSB	2528	(4068)
EME	K2UYH (FN20QG)-VK5MC (QF02EJ)	81-12-06	CW	10,562	(16995)
Tropo	WB3CZG (FN21AX)-KD5RO (EM13PA)	86-11-29	CW	1287	(2070)
2304 MHz					
EME	W3IWI/8 (FM08CK)-ZL2AQE (RE78JS)	87-10-18	CW	8658	(13931)
Tropo	KD5RO (EM13PA)-W8YIO (EN82BE)	86-11-29	CW	940	(1513)
3456 MHz					
Tropo	WA5TNY/5 (EM11AU)-WB5LUA/5 (EM24UQ)	86-10-19	CW	288	(464)
EME	W7CNK/5 (EM15FI)-K0KE/0 (DM79NO)	87-04-12	CW	498	(802)
5760 MHz					
Tropo	K5PJR (EM26OP)-W5UGO/0 (EN00PH)	87-07-04	CW	332	(535)
EME	WA5TNY (EM12KV)-W7CNK/5 (EM15FI)	87-04-24	CW	174	(279)
10.368 GHz					
Tropo	N6GN/6 (CM89PX)-W6SFH/6 (DM04MS)	87-07-19	CW	414	(666)
24.192 GHz					
LOS	WA3RMX/7 (CN93IQ)-WB7UNU/7 (CN95DH)	86-08-23	SSB	116	(186)
47.040 GHz					
LOS	WA3RMX (CN85PL)-WB7UNU/W7TYR (CN85NH)	87-03-08	SSB	13.9	(22.4)
76-149 GHz	None reported				
474 THz					
LOS	K6MEP (DM04IO)-WA6EJO (DM04KT)	79-06-09	LASER	15	(24)

Note 1. The records are listed alphabetically by mode. Ducting is suspected where the path is mostly over water. No efforts are made to separate out ducting on overland paths so they're grouped under tropo.

Note 2. The information within the parentheses ( ) following the callsign is the grid square locator.

Note 3. Distances have been calculated assuming a spherical earth model using the actual latitude and longitude rather than grid square centers which are less accurate.

Note 4. Six-meters records, excepting EME, were left off since the primary propagation mode is often hard to distinguish. Also long-path QSOs have been reported during solar cycles 19 and 21 which exceed approximately 12,430 miles.

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Table 2. Worldwide Claimed VHF/UHF/SHF Terrestrial DX Records (notes 1 & 2)

Frequency	Record Holders	Date	Mode	DX Miles	(km)
50 MHz	Note 3				
70 MHz	GW4ASR/P (IO82JG)-5B4CY (KM64MR)	81-06-07	Es	2153	(3465)
144 MHz	I4EAT (JN54VG)-ZS3B (JG73OI)	79-03-30	TE	4884	(7860)
220 MHz	KP4EOR (FK68XM)-LU7DJZ (GF05RJ)	83-03-09	TE	3670	(5906)
432 MHz	KD6R (DM13NI)-KH6IAA/P (BK29GO)	80-07-28	Duct	2550	(4103)
903 MHz	W2PGC (FN02OR)-K3SIW/9 (EN52WA)	86-12-24	Tropo	478	(769)
1296 MHz	KH6HME (BK29GO)-WB6NMT (DM12KU)	86-08-13	Duct	2528	(4068)
2304 MHz	VK5QR (PF95HD)-VK6WG/P (OF85WA)	78-02-17	Duct	1170	(1883)
3456 MHz	VK5QR (PF95HD)-VK6WG (OF85WA)	86-01-25	Duct	1171	(1885)
5670 MHz	G3ZEZ (JO01MS)-SM6HYG (JO58RG)	83-07-12	Duct	610	(981)
10.3 GHz	I0SNY/EA9 (IM75IV)-I0YLI/IE9 (JM68NR)	83-07-08	Duct	1032	(1660)
24 GHz	I0SNY/IC8 (JN60WR)-I8YZO/8 (JM78WE)	84-08-11	LOS	206	(331)
47 GHz	HB9AGE/P (JN36FS)-HB9MIN/9 (JN36SX)	87-06-06	LOS	53.5	(86)
75 GHz	HB9AGE/P (JN37RD)-HB9MIN/P (JN37RD)	85-12-30	LOS	0.3	(0.5)
474 THz	K6MEP (DM04IO)-WA6EJO (DM04KT)	79-06-09	LOS	15	(24)

Notes:
1. The information within the parentheses ( ) after the callsign is the grid square locator.
2. Distances have been calculated assuming a spherical earth model. The actual latitude and longitude are used rather than grid square centers which are less accurate.
3. Six meters has been left blank on this listing because long-path QSOs (those exceeding approximately 12,430 miles) have been reported during solar cycles 19 and 21.

Table 3. Worldwide Claimed VHF/UHF/SHF EME DX Records (notes 1 & 2)

Frequency	Record Holders	Date	Mode	DX Miles	(km)
50 MHz	WA4NJP (EM84DG)-KH6HI (BL01XH)	88-02-15	CW	4530	(7289)
144 MHz	K6MYC/KH6 (BK29AO)-ZS6ALE (KG43RC)	83-02-18	CW	12,091	(19455)
220 MHz	K1WHS (FN43MK)-KH6BFZ (BL11CJ)	83-11-17	CW	5058	(8139)
432 MHz	F9FT (JO29AG)-ZL3AAD (RE66GR)	80-04-18	CW	11,679	(18793)
902 MHz	K5JL (EM15DQ)-WB5LUA (EM13QC)	88-02-07	CW	187	(301)
1296 MHz	PA0SSB (JO11WI)-ZL3AAD (RE66GR)	83-06-13	CW	11,595	(18657)
2304 MHz	W3IWI/8 (FM08CK)-ZL2AQE (RE78JS)	87-10-18	CW	8658	(13931)
3456 MHz	W7CNK/5 (EM15FI)-K0KE (DM79NO)	87-04-06	CW	498	(802)
5670 MHz	WA5TNY (EM12KV)-W7CNK/5 (EM15FI)	87-04-24	CW	174	(279)
10,000 MHz and above: None reported					

Notes:
1. The information within the parentheses ( ) following the callsigns is the grid square locator.
2. The distances shown have been calculated assuming a spherical earth model. The actual latitudes and longitude are used rather than grid square centers which are less accurate.

still in effect in Colorado, Wyoming, White Sands Missile Range, and Region 3 areas. Operators in these restricted areas who have tried to obtain permission from the FCC have been unable to do so. Canadian Amateurs need special permission from DOC to use CW or SSB on this band. (It is presently designated as fm only!)

solar cycle update

Probably one of the hottest discussions on the hf and 6-meter bands these days is "when will the next solar cycle peak?" Near the sunspot peak there is a chance that F2 propagation will be possible on 6 meters. News of that peak is starting to come in. The smoothed sunspot count of Cycle 21 peaked at 164.5 in December of 1979 and ended when it bottomed out at 12.3 sunspots in September 1986. The new cycle, 22, has definitely begun and no one knows for sure how high the peak will be or when it will occur.

Improved methods of forecasting like the Sargent/Ohl[7] were very close in predicting the peak of cycle 21. Based on available data and using this prediction method[8], it now appears that cycle 22 will peak at a smoothed level of 118.6 sunspots in mid-1991. **Figure 1** shows this early data along with the final data on cycle 21.

The predicted peak of cycle 22 shows that it will be very flat and should stay above 100 sunspots from about July 1989 through June 1992. Because this cycle started statistically

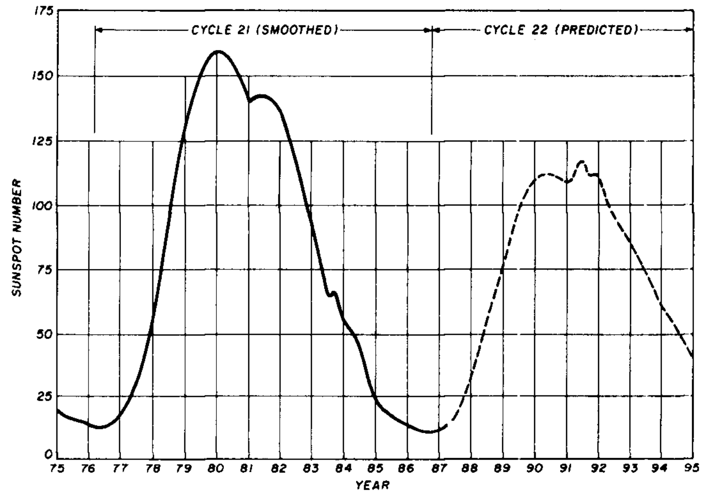

fig. 1. This graph shows the smoothed sunspot numbers for solar cycle 21 as well as the forcasted numbers for cycle 22, per reference 8.

earlier than other cycles and rose abruptly, we will have to wait at least another year or so to see what if any modifications will occur.

This information is not very promising for 6-meter Amateurs as it usually takes a sunspot count above 150 to yield good F2 openings. However, minor sunspot peaks often occur during a cycle, albeit of short duration. No 6-meter operator active during the last solar cycle will ever forget the solar peaks in late 1979 that rivaled those of all previously recorded solar cycles.

The equivalent short-term sunspot number can be predicted using the solar flux measured at Ottawa on 10.7 cm. The value is updated daily and broadcast at 18 minutes after each hour on radio station WWV. Using the equation shown in **reference 2**, I have prepared **fig. 2** which can be used to determine the equivalent sunspot number on any day. Remember also that the ionosphere usually has to be "pumped up" for four or five consecutive days to yield good long-haul F2 propagation.

The SOA in equipment, antennas, and propagation forecasting has greatly improved in recent years. Predictions of the MUF are now possible with improved accuracy using personal computer programs like MINIMUF[8] in conjunction with the sunspot number per **fig. 2**.

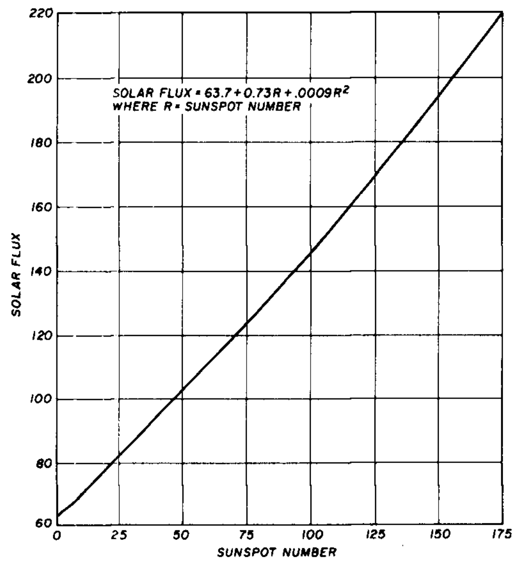

fig. 2. This figure shows the correlation between solar flux and sunspot numbers.

This information, along with increased 6-meter interest and improved operating methods (more on this shortly), as well as recent relaxations in licensing restrictions in western Europe and North Africa, means that there will be many more regions and DXCC countries represented during cycle 22. Let's hope it's a great cycle for 6-meter operators. Stay tuned!

backscatter

The backscatter form of propagation described in **reference 1** is basically a form of reflection and indicates that a highly ionized region is present. Operators detecting this phenomenon can often work DX by aiming their antennas in the direction of the ionized region and "backscattering" their signals.

Backscatter also indicates that the MUF is very high; it was well used during solar cycle 21 to indicate the presence of an opening. Often western United States stations could work Hawaii while eastern stations could work Europe either by backscatter or by knowing that there was a high degree of probability of an opening in progress.

Look for backscatter especially over the Atlantic and Pacific Oceans. It can also be observed on 10 meters to indicate a possible 6-meter opening. Backscatter will become much more evident as the sunspots increase and is usable up through 6 meters.

ionospheric scatter

Ionospheric scatter was also described in **reference 1.** It is a form of "forward scatter" linked to the time

Table 4. VHF/UHF/SHF Propagation Record Verification Form.
(Please return to Joe Reisert, W1JR, 17 Mansfield Drive, Chelmsford, MA 01824)

Band:________________Propagation Mode:________________Date of record (UTC):________________

Time of record (UTC):________________DX (miles)________________(km):________________

Station 1

Call:________________

Name:________________

QTH for this QSO:________________

Lat*:________________Long*:________________

Grid Locator (6 digit):________________

Elevation ASL (feet):________________(meters):________________

Location description:________________

Antenna type:________________

Estimated gain (dBi):________________

TX freq:________________

TX power:________________

Feedline loss:________________

Modulation type:________________

RX freq:________________

RX type:________________

RX desc:________________

Feedline loss:________________

Noise figure:________________

RX bandwidth:________________

Rcvd signal to noise ratio:________________

Other equipment description:________________

Station 2

Call:________________

Name:________________

QTH for this QSO:________________

Lat*:________________Long*:________________

Grid Locator (6 digit):________________

Elevation ASL (feet):________________(meters):________________

Location description:________________

Antenna type:________________

Estimated gain (dBi):________________

TX freq:________________

TX power:________________

Feedline loss:________________

Modulation type:________________

RX freq:________________

RX type:________________

RX desc:________________

Feedline loss:________________

Noise figure:________________

RX bandwidth:________________

Rcvd signal to noise ratio:________________

Other equipment description:________________

Other comments, weather conditions etc:________________

The information submitted above is correct to the best of my knowledge.

Submitted by:________________Call:________________Home QTH:________________

________________Phone number (AC________)________________

Record received and verified by:________________Date:________________

*Please list latitudes and longitudes in degrees, minutes, and seconds.

of day (typically peaking broadly around noon local time) and to high sunspot activity. Ionospheric scatter can be used on 6 and 2 meters.

This form of propagation is used extensively in commercial service but seems to have been almost totally ignored by Amateurs. It does require reasonable antenna gain and high power, but is within the reach of many well-equipped 2-meter Amateur stations, especially those with EME or marginal EME capability.

As the sunspots increase, so will the possibilities of forward scatter. This represents an interesting challenge for Amateurs and is a good way to increase their grid square count in the 800-1300 mile region.

TE (transequatorial) scatter

Like forward and back scatter, TE propagation is a good mode for long DX, especially on 6 meters. It is best observed across the equator on more or less directly north-south paths with typical distances of 3 to 6,000 miles.

TE propagation is most often observed in the late afternoon and early evening for several weeks around the equinoxes. During the peak of the solar cycle around this time, highly ionized "patches" are often present approximately 10-20 degrees north and south of the "geomagnetic" equator.

Unfortunately, the geomagnetic equator is very far south in the North and South American sector. This limits North American TE propagation mainly to stations in the Caribbean and the extreme southern portions of the United States. Don't let this discourage you; there are always new propagation modes and isolated openings to explore.

equatorial FAI (Field Aligned Irregulatities)

Equatorial FAI was first discovered in 1977.[10] It is still not fully understood and often referred to as TE propagation (see **references 11** and **12**). Like TE scatter, equatorial FAI depends on highly ionized patches that are typically located 10-15 degrees north and south of the geomagnetic equator at the same dates and times discussed under TE propagation above. However, the DX is slightly less. The most favored locations are paths from southern Europe to South Africa, Japan to Australia, and the Caribbean to southern South America.

While TE scatter is generally limited to below 100 MHz, equatorial FAI has been known to extend higher in frequency. Contacts as high as 220 MHz have been confirmed as shown in **tables 1** and **2**. Although some one-way 432 reports have been reported, I have been unable to document any two-way contacts above 220 MHz. Maybe during the peak years of solar cycle 22 the frequency barrier will be broken and two-way 432 MHz contacts will be completed. Any takers?

midlatitude FAI

When **reference 10** was written, it was speculated that FAI propagation would be possible in mid-northern latitudes. It didn't take long before this became a reality on 144 MHz.[13]

Midlatitude FAI propagation has many similarities to auroral propagation; both stations must be south of the ionized region and aim their antennas several degrees north of the great circle path. This type of propagation most often occurs in the evenings during the summer, especially on days when there has been sporadic E propagation on 6 meters. As shown in **table 1**, it has been successfully used out to a distance of just over 1200 miles on 2 meters. Until recently FAI has been slow to take hold, despite the fact that it should be usable up through 220 MHz.[13]

This is all changing now — contacts were reported during the summer of 1987 in the southern United States and a wide region of Europe (**reference 14** through **16**). In fact, well over 500 European contacts were reported using midlatitude FAI propagation during the summer period of 1986 alone! (See **reference 15**.)

As observed in Europe, the scatter region tends to be at the same height as sporadic E, typically 70 miles. These regions resemble aurora propagation; unlike the relatively small (1-2 mile thick) sporadic E clouds, they have large volume areas.

Midlatitude FAI signals tend to have rapid fading. They have been observed over several European locations but mostly along the 45-55 degree north latitude lines following the contours shown in **reference 13**. Those who can elevate their antennas have a greater possibility of success.

Midlatitude FAI propagation offers a great challenge to VHF Amateurs, especially in North America. This mode of propagation should be usable up through 220 MHz throughout the contiguous 48 United States. All it takes is some patience and a surge in activity. Who will be the first to report a 220-MHz midlatitude FAI QSO? It's there for the asking!

summary

This month I've given you a status report on the latest DX records on the VHF/UHF/microwave bands. We've also discussed the latest prognosis for propagation using the solar cycle 22 peak and some scatter modes. Next month's column will update other propagation modes. Until then, you can read the references cited.

new DX records

This has been a good month for new VHF/UHF DX records. First off, the 6-meter EME record has been extended. On February 15, 1988 between 1800-1845 UTC, Ray Rector, WA4NJP, Gillsville, Georgia (EM84DG) completed a two-way EME contact with Bert Ingalls, KH6HI, Ewa Beach, Hawaii (BL01XH), on 50.008 MHz using 1-minute sequencing. The distance was approximately 4530 miles (7289 km). Ray was using 1500 watts and Bert was running 1000. Both stations were using quads of four eight-element Yagis on 35-38 foot booms. Congratulations to Ray and Bert — 6 meters is now buzzing with EME activity.

Last month we reported the first ever 33-cm (902 MHz) EME QSO. That record didn't last very long! On February 7, 1988 at 0500 UTC, Jay Lieb-

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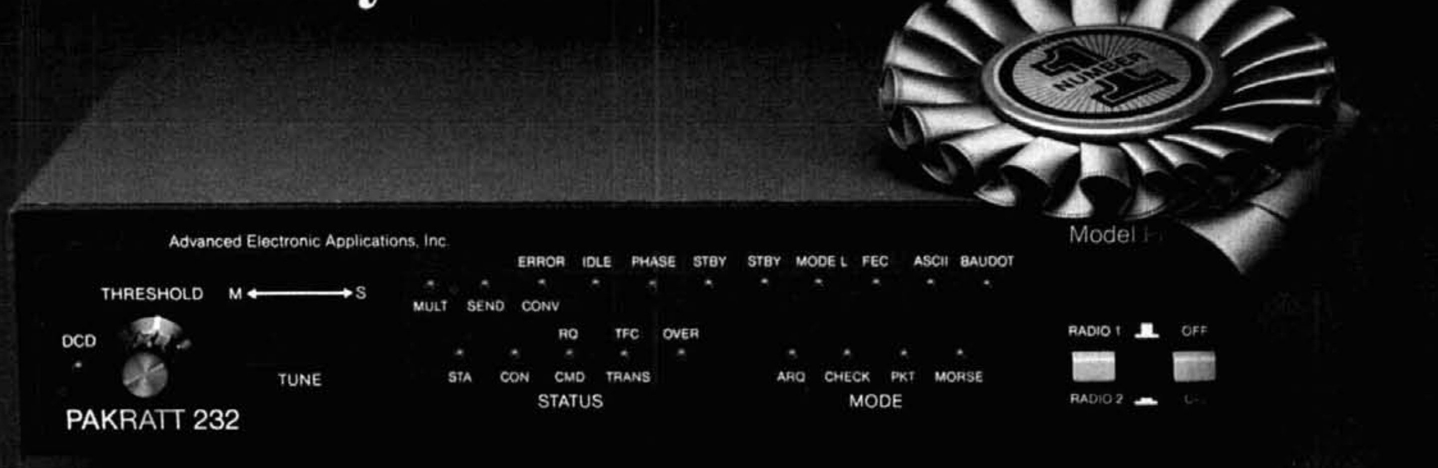

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Reader Service CHECK—OFF Page 98 ✓ 146

mann, K5JL, Piedmont, Oklahoma (EM15DQ) completed a 33-cm EME QSO with Al Ward, WB5LUA, McKinney, Texas (EM13QC) over a distance of approximately 187 miles (301 km). Both stations were running approximately 150 watts and 24-28 foot dishes. Congratulations, Jay and Al. It looks as if 33-cm EME activity is just about to take off. Both records just discussed are included in **tables 1** and **3**.

Finally (although not yet a DX record), during February 1988 Rick Fogle, Grapevine, Texas (EM12KV) has been heard by Lucky Whitaker, W7CNK, Oklahoma City, Oklahoma (EM15FI) via 3-cm (10,368.1 MHz) EME. Likewise, Lucky has heard Rick via the same path. Rick uses a 10-foot dish and Lucky a 16 footer. Both have their preamplifiers and power amplifiers mounted right at the feed. Unfortunately, they have only one high-power (10-15 watt) TWT amplifier between them which they mail back and forth. Because of this, they can't complete what is considered a conventional two-way QSO (use of two complete sets of gear all used during one operating session). Efforts are underway to get a second power amplifier. Good luck to Rick and Lucky as well as the other 3-cm operators who are also trying to conquer this elusive band. It seems that one of the last EME frontiers is about to be conquered.

important VHF/UHF events

June 4	*EME perigee*
June 7	*Predicted peak of the daytime Arietids meteor shower at 0150 UTC*
June 9	*Predicted peak of the Zeta Perseids meteor shower at 1020 UTC*
June 11-13	*ARRL June VHF QSO Party*
June 14	*New moon*
June 18-19	*SMIRK (Six Meter International Radio Klub) Party Contest (contact KAØNNO)*
June 21	*± 1 month. Peak of midlatitude Sporadic E propagation*
July 1	*± 1 month. Look for United States to Europe openings on 6 meters*
July 2	*EME perigee*
July 13	*New moon*
July 16-17	*CQ Magazine VHF WPX Contest*
July 20	*± 3 weeks. Look for 2-meter Sporadic E openings*
July 21-24	*Central States VHF Conference, Lincoln, Nebraska, NEB (contact WDØDGF)*
July 28	*Predicted peak of the Delta Aquarids meteor shower at 2100 UTC*
July 30	*EME perigee*

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ham radio

the Quad antenna: part 2,
circular and octagonal loops

Analysis shows good performance with similar data

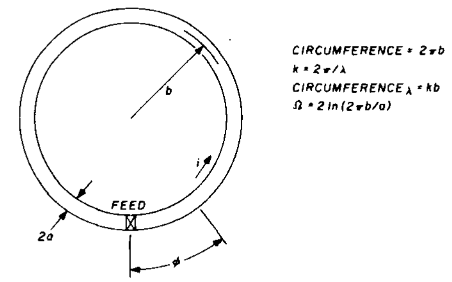

fig. 1. A circular loop can be described by two quantities, the radius of the conductor (a) and the radius of the loop (b). For work, it is convenient to use two derived quantities, a conductor thickness factor (omega) and the circumference in wavelengths at the operating frequency (kb).

Although there is a contradiction in terms, it is convenient to consider the circular loop, and arrays built of loops, as the first members of the Quad family. One reason is that all other versions can be regarded as departures from the "ideal" circular figure. To the extent this is true, the performance of circular-loop antennas is thus representative of the performance of the entire family.

Another reason is that the theoretical anaylsis of the circular loop is far more advanced than for the other shapes. Extensive tables of calculated characteristics have been published, some with comparisons of measured performance. In contrast, while there are theories of square Quad loop and array performance, their complexity makes them impractical for calculation, even on mainframe computers.

theoretical basis of circular-loop analysis

As shown in **fig. 1**, only two quantities are necessary to specify the circular-loop antenna: the conductor size, usually given as its radius; and the loop size, also described by radius. It is often more convenient to use two derived descriptive quantities in theoretical discussion. The first is the normalized circumference of the loop at the specific frequency of interest given by the quantity kb, where k is defined as

$$k = 2\pi/\lambda,$$

By R.P. Haviland, W4MB, 1035 Green Acres Circle North, Daytona Beach, Florida 32019

λ being the wavelength. The quantity kb is therefore the circumference of the loop in wavelengths. Conductor size is usually given by the relationship

$$\Omega = 2ln\left(\frac{2\pi b}{a}\right)$$

where ln is the natural logarithm, equal to 2.3 times the more common $\log_{10}$ value. The value of omega is given in **fig. 2** as a function of the ratio 2πb/a, or loop circumference to conductor radius. Values less than 10 represent very large conductor diameters, and those over 20 very small conductors. High-frequency antennas will usually have values in the range 20-25, and self-supporting ultrahigh frequency antennas values of 10-15.

The loop is assumed to be fed at one point, usually taken as the angle reference. This induces a current, *i*, in the loop at angle zero which, in turn, creates a field at the point designated by R,01, for example. The total field at this point is the sum of the fields produced by all points on the loop.

The field components also induce current flow in the loop. When equilibrium is reached (after a few rf cycles) the field close to the conductor must lie only at right angles to it. (If there had been a tangential component, a change in the current would have been induced, so equilibrium would not yet have been attained.) This observation plus standard field equations give the conditions for calculating current distribution, and therefore the drive impedance and radiation pattern.

While the concept is relatively simple, the mathematical operations are difficult. See the references, especially Storer,[1] for details. For our purposes it is sufficient to note that the current distribution is given by:

$$I(\emptyset) = \frac{V}{(j \cdot \pi \cdot 377)}\left[\frac{1}{A_o} + 2 \cdot sum\frac{Cos(n \cdot \emptyset)}{A_n}\right]$$

where the sum is for all values of N from 1 to infinity. This result was derived by Hallen.[2]

This equation is simple, but its evaluation is complex. The quantities A involve series for which exact solutions are unknown. Even approximate solutions require further assumptions, two being that the conductor diameter is small compared to loop diameter and to operating wavelength. This restriction is satisfied by practical antennas.

Additionally, the infinite series in the equation tends

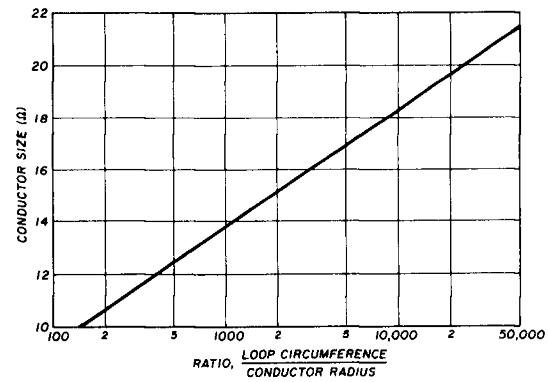

fig. 2. Values of the thickness factor (omega) as a function of the ratio of loop to conductor radius (or diameter). Practical self-supporting loops need an omega around 10-12 to have sufficient strength. Wire cage elements may be used to secure low omega factors at low frequencies.

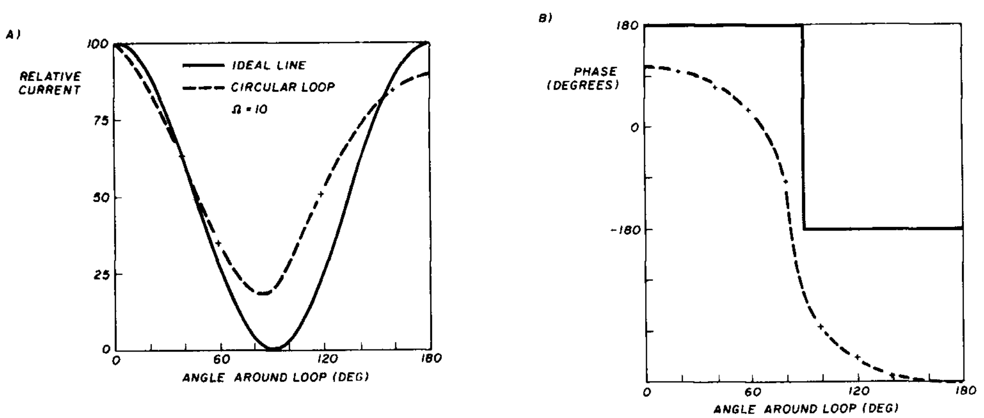

fig. 3. Current magnitude and phase on a large conductor circular loop one wavelength in circumference as derived by Storer.[3,1] The loop is below resonance. The current on a conductor of essentially 0 radius is shown for comparison. Derived from transmission line theory, this is the first approximation to the current on any loop of the same circumference.

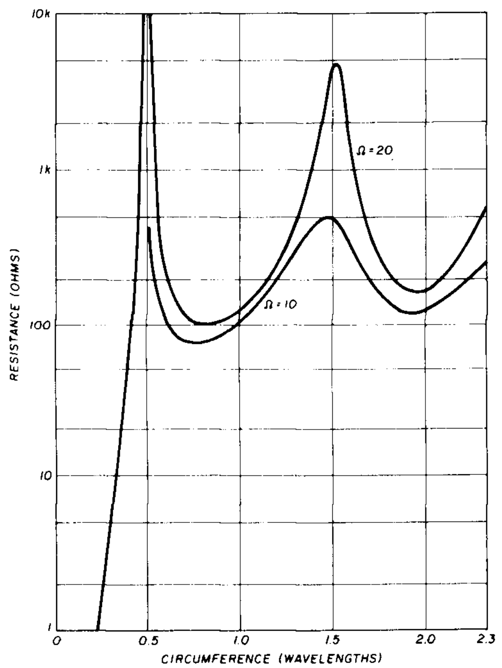

fig. 4A. Feedpoint or self-resistance of one-wavelength circumference loops — one with a large conductor radius, the other small. The drive resistance is sensitive to conductor size around the points of parallel resonance, at 0.5, 1.5, 2.5, etc. wavelengths circumference. These curves may be used as approximations for any loop of equal circumference and radius.

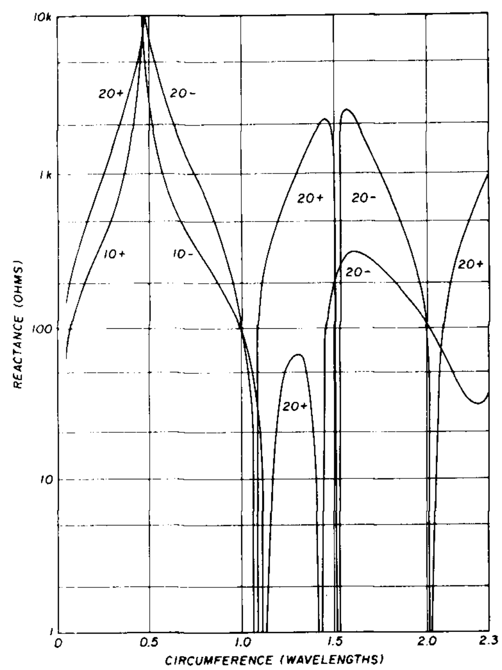

fig. 4B. Feedpoint or self-reactance of one-wavelength circumference loops. High reactance denotes the parallel resonances, low reactance the series resonances. Both series and parallel resonances are sensitive to conductor size. Series resonance is generally at wavelengths greater than integral values. Parallel resonances are at wavelengths less than odd multiples of 0.5 wavelength.

to become divergent when the number of terms is reduced for reasonable scale of calculation. Storer[3.1] developed a method of calculation by keeping the first four terms of the series and replacing the remainder by an integral. He also published a set of ten curves giving the real and imaginary components of the series. With these, evaluation of the current distribution and drive impedance is reduced to some simple (but tedious) curve measuring and complex-number algebra. Unfortunately, the curves cannot be reduced to simple equation form, so this can't be avoided.

Rather than presenting these curves and usage details here, I will give only the results of examination of some specific loop designs. Subsequent analyses have given a table of drive impedances, which is more accurate for most work. These values are covered later.

For values outside the range given here, or to obtain the current distribution, you will need the Storer curves. (The **reference 3** version is best, and is available at reasonable cost from Cruft Laboratory Library, Harvard University, Cambridge, Massachusetts.)

A simpler solution for small circular loops is available.[4] Here "small" covers the range from 0 to 0.3 wavelength circumference. This restriction allows

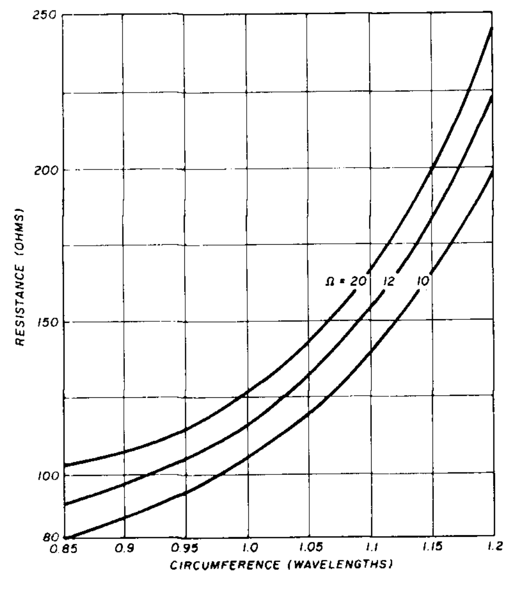

fig. 5A. Feedpoint or self-resistance of three loops around one-wavelength circumference.

simplification of the general equation above, with only three terms giving adequate accuracy. Approximate equations for each are given, and are easily handled by small computers. Note that the reactance for these small antennas is determined almost entirely by the loop circumference, with almost no effect on conductor size. In contrast, the input resistance varies with both.

current distribution on a circular loop

Figure 3 shows the magnitude of the current on a one-wavelength circular loop as calculated by Storer, together with the current magnitude on an ideal shorted transmission line made from one wavelength of conductor. **Figure 3B** shows the phase with respect to the driving voltage for both cases.

A number of important factors concerning the entire Quad family show on these curves. The first is that a one-wavelength loop is not resonant, as indicated by the fact that the angle of the drive impedance is not zero. Since the loop appears as a capacitance, it is below resonance. Unlike dipoles, loops must be longer than a wavelength to be resonant.

A second factor is that there is no point on the loop where the current goes to zero. Associated with this is the fact that the current at 180 degrees from the feedpoint is appreciably less than at the driving point. Similarly, these two currents are not 180 degrees out of phase, but somewhat more. The point of phase reversal is not at 90 degrees to the line of symmetry through the feedpoint.

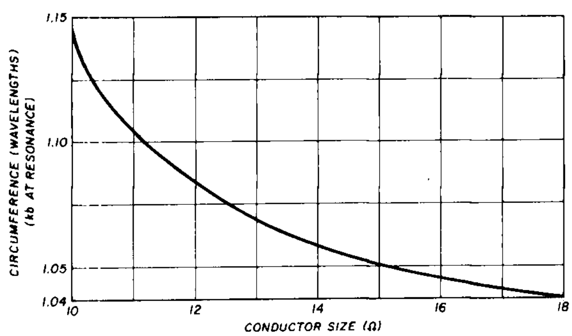

fig. 6. Resonant frequency of a circular loop derived from *fig. 5B*, as the frequency of 0 reactance. A simple linear relationship of reactance versus wavelength (or frequency) can be developed using two figures. However, the curves are usually less time consuming to use.

One reason for the differences between the loop and the ideal shorted line is the greater separation of the sides. Currents are not constrained to be equal because of tight coupling, as in the ideal shorted line. Further, power is being radiated, causing a reduction in current when moving away from the feedpoint. (In a practical antenna, the current differences would be somewhat greater, as the analysis assumes zero ohms loss.)

These current curves show that the usual evaluation of a Quad — two separated dipoles with ends bent to touch — can't fully describe the performance. This simple concept is useful in verbal descriptions, and can be a valuable tool in approximate analysis. But it must be remembered that numerical results are probably in error by a factor *at least as large as the current error*, or at least 20 percent. The effect of the error should be smallest for pattern calculations and drive resistance, but is likely to be sizeable for drive reactance and resonant frequency.

These detail current calculations are for loops with relatively large conductors, $\Omega = 10$. Other studies, plus the tables presented later, show that the current magnitude and angle move progressively toward the curve for the ideal transmission line as the conductor radius becomes smaller. This means that the two-dipole approximation is likely to be better at high frequency than at ultrahigh frequency because omega is large for practical conductor sizes. The current distribution for some other loops is also given in **references 1, 3, 5,** and **6**.

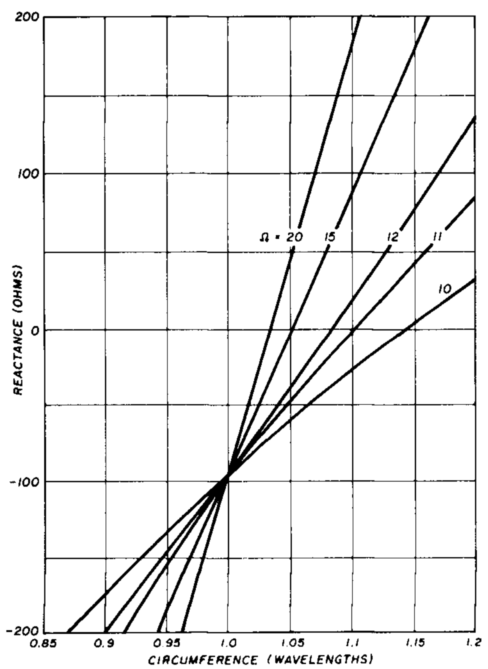

fig. 5B. Feedpoint or self-reactance of five loops around one-wavelength circumference. The reactance for all conductor sizes is equal to −95 ohms for a circumference of one wavelength. This may be compared with dipole reactance which shows the same characteristic, with a reactance of 35 ohms at 0.94 wavelength as the common point.

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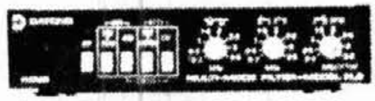

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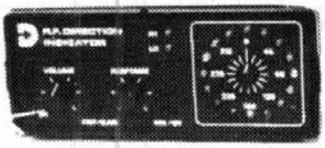

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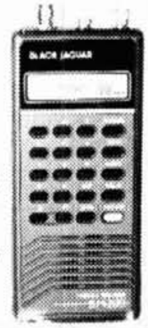

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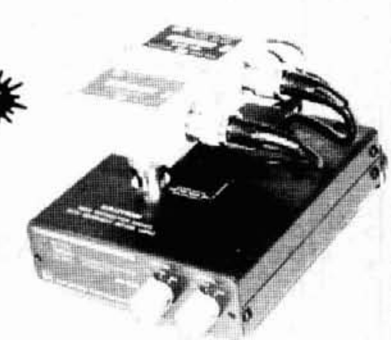

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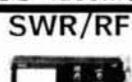

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As a first approximation, the current distribution can always be assumed to be that of an ideal transmission line of the same conductor length. A second approximation can be sketched by "rounding all sharp corners," and decreasing the current away from the feedpoint. Greater accuracy will require tedious evaluation using Storer's curves or the MININEC technique (to be discussed).

drive-point impedance

Get the drive-point impedance by dividing the drive-point voltage by the drive current calculated above. Storer[3,1] gives tables of this impedance for loops from 0.05 to 2.5 wavelengths in circumference, and for omega values from 8 to 12 (large diameter conductors).

In considering the general problem of loop antennas, Wu[7] developed another method for solving Hallen's equation. This gives the same values for resistance components as Storer, but the reactance values are quite different and agree more closely with measurements on real antennas. King, Harrison, and Tingley[8,9] have used Wu's theory to calculate loop-drive admittances for sizes from 0.05 to 2.5 wavelengths and for omega from 10 to 20. For those who have forgotten, or never had occasion to work with admittances,

$$z = R + jX$$
$$y = 1/Z$$
$$y = g + jb$$

(See also **reference 10.**)

Figures 4A and **4B** show feed resistance (R) and reactance (X), respectively, for loops from 0.05 to 2.5 wavelengths circumference. The two values of omega shown, 10 and 20, are representative of very high frequency and high-frequency loops.

Over this range of loop diameters, three high-impedance or parallel resonance points are noted, corresponding to 0.5, 1.5, and 2.5 wavelengths circumference. For $\Omega = 20$, the low-impedance, zero-reactance points correspond to serial resonances at about 1.0 and 2.0 wavelengths circumference. For $\Omega = 10$, there is also a serial resonance at 1.0 wavelength circumference. There is no true serial resonance for a circumference of 2.0 wavelengths. Instead, the reactance becomes low and remains low and capacitive. This is the case for all values of omega less than 11, and for all but the first serial resonance.

The resistance at the serial resonance point changes markedly with the value of omega. For omega that is large (20 or so), the resistance also varies markedly with loop diameter. But for $\Omega = 10$, the resistance change is much smaller. (For $\Omega = 8$, the change is less than 2:1 for any circumference greater than 0.6 wavelength.)

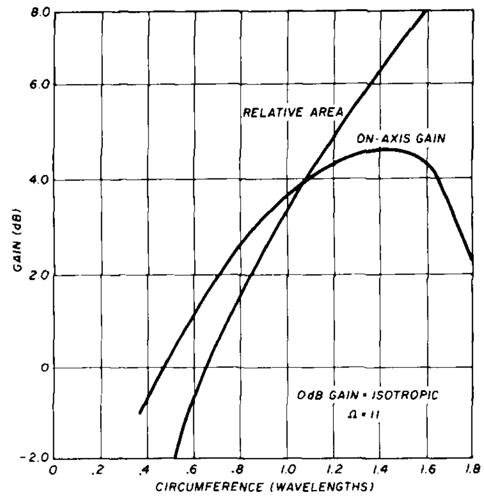


fig. 7. Gain of a circular loop in dB above isotropic, as derived by Ito el al.[11] Maximum gain occurs for a loop circumference of 1.4 wavelengths. Gain is less for smaller sizes due to less radiating area. On-axis gain, the value shown, is less at larger sizes because of lobe splitting. However, the gain of the main lobe increases as circumference increases. The associated curve shows the relative area of the loop.

These characteristics mean that thick loops are inherently broadband antennas and relatively easy to match. However, the pattern changes described in part 1 affect the desirability of this broadband operation, as discussed later.

Figures 5A and **5B** show the feed resistance and reactance for frequencies of greatest interest (those close to the first series resonance) for loops from 0.8 to 1.2 wavelengths circumference. **Figure 6** is derived from **fig. 5B**, and shows the resonant frequency as a function of conductor size. These three curves give the information needed to design practical loop antennas and arrays and to calculate their performance. Their use is covered further on.

gain of loop antennas

In considering loop gain, it was noted that gain should increase as loop size increases and that there are pattern changes as size increases. Specifically, gain on the axis of symmetry becomes zero for all loops with 2, 3, or more wavelengths in conductor length.

These two effects are shown in the calculated gain curve of **fig. 7** (see **reference 11**). For a circumference of 1 wavelength, the on-axis gain is 3.4 dB above isotropic, or about 1.4 dB above a dipole. (Based on measurements, Lindsay[12] quotes approximately 4.0 dB, and Appel-Hansen[13] quotes 3.4 dB above isotropic.)

Figure 7 shows that larger loops are superior from the viewpoint of gain. A calculated gain value of 4.5 dB occurs for a loop circumference of 1.5 wavelengths. Above this circumference, the gain decreases, as the on-axis lobe splits. Remember, there is no radiation on axis for a circumference of 2 wavelengths.

In smaller sizes, the on-axis gain decreases as the circumference becomes less than a wavelength, and as the pattern changes toward the doughnut pattern of a small loop. The gain is about equal to that of a dipole at a circumference of about 0.65 wavelength, and about 1.5 dB poorer than that of a dipole for a circumference of 0.5 wavelength.

This brings up two important points. First, from the view of gain, the circular loop should be designed to operate away from resonance. Considering the factors of gain, lobe shape, and feed impedance, Ito et al.[11] recommend a design circumference of 1.2 wavelengths, for a gain of 4.2 dB.

Second, a loop has good gain performance over a wide range of frequences. For example, a 14-MHz loop would be slightly better than a dipole on 10 MHz, and about 1.5 dB poorer on 7 MHz. The gain would be near maximum on 21 MHz, better than a dipole on 18 MHz, and about as good on 24 MHz. The on-axis gain would be poor on 28 MHz, but there would be usable radiation in two lobes.

The acceptability of operation away from resonance is affected by the difficulty encountered in feeding the loop. From **fig. 4**, feeding a large loop on high frequency does not appear to be an unusual problem. The feed resistance would increase, but a matching section or transformer is needed. There would be an inductive component, easily compensated by a stub. Depending on conductor size, a very high frequency loop might require only a matching transformer.

Wideband operation would be a greater problem.

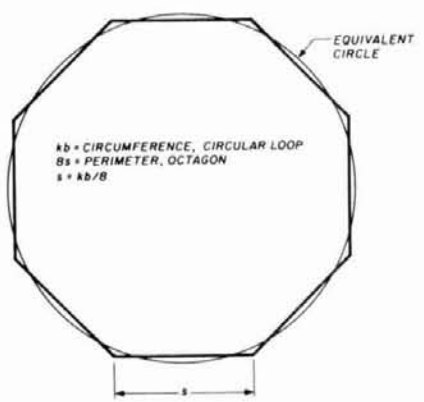


fig. 8. Relative geometry of a circle and an octagon. The difference in area for the same conductor length is very small.

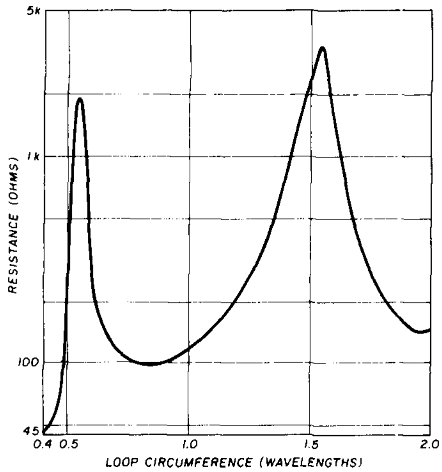
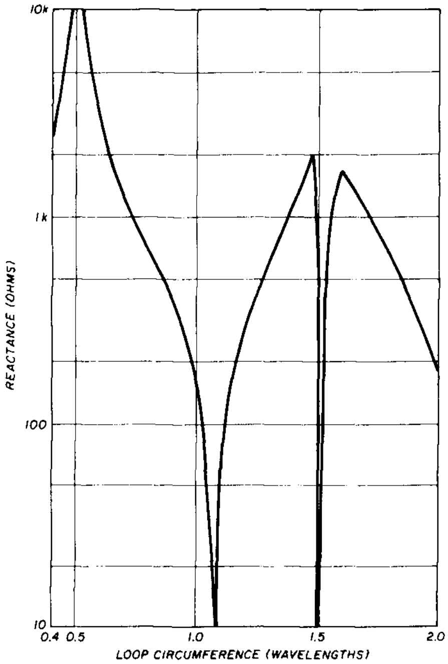


fig. 9. (A) Feed or self-resistance of an octagonal loop for Ω = 17.9, calculated by MININEC 3.[14] Values are within a few percent of those for the corresponding circular loop. (B) Feed or self-reactance of an octagonal loop. Differences between these and circular loop values are somewhat geater than for resistance. They are due to different area, and the geometry of current interaction. It is likely that there are also differences due to the calculation approximations, since the expected calculation error may reach 10 percent.



You'd need a complex matching section at the antenna, or your transmission line would have appreciable VSWR. The combination of open-wire line and a wide-range "Match Box" would give excellent performance. This is especially interesting as the basis of a 3.5/7/10-MHz fixed-loop design.

I have made use of this wideband capability on many occasions. A 14-MHz loop is regularly used on bands from 7 to 28 MHz. Much of my 10, 18, and 24-MHz experimental work (KK2XJM) used this antenna. Other than Teflon™ insulated 75-ohm feedline and a pi-section tuner, no special precautions were taken.

circular-loop radiation patterns

Published data on radiation patterns of loop antennas varies from sparse to nonexistent. The equations needed to calculate the currents and patterns are solvable on a small computer only with a lot of programming and computing time. It has been necessary to approximate further to develop the pattern data which follows.

the MININEC antenna program

The chosen calculation technique (sometimes called "the method of moments") is the public domain program "MININEC," from Logan and Rockway,[14] currently in its third version. It uses essentially the same initial assumptions as the Hallen[2] approach above. But instead of applying the geometry exactly and then making simplifying assumptions, assume that the radiator is composed of a series of straight-line sections carrying constant current.

A solution's accuracy increases as the number of sections for a given geometry is increased. While the complexity of the solution is not greatly affected, the time required for the solution increases as the square of the number of segments. An IBM PC may take several hours to run the program; a very small computer like the Commodore 64 may need 12 to 24. Even so, this approach is the best generally available, and is quite practical.

The original MININEC[15] was written for the Apple computer. I have translated it for the C-64 and the Amiga; KA4WDK has done the same for the PC. The third version was written for the PC, and I translated it for the Amiga; it also runs on the Macintosh.

MININEC originators used various conditions to examine calculation accuracy, including analysis of loops. One series used a ten-sided polygon approximation of a circle, with two current segments per side plus three for the feed side. Agreement with theory was to within 10 percent.

A second series approximated the circle using the circumscribed polygon. This shows good agreement (±6 percent) for susceptance, down to four sides. Equally good agreement for conductance required 16 sides. With 22 sides, agreement was within 6 percent for sizes from 0.1 to 2.0 wavelengths. MININEC solutions are found to be unreliable for circumferences less than about 0.01 wavelength.

The inscribed polygon approximation is not an ideal check of solution accuracy as the number of segments decreases. It introduces two added factors which change the results. One is the loss of area. (For eight sides this amounts to about 3 percent.) This reduces the gain by about the same amount. More important, the total conductor length decreases as the number of segments decreases, by about the same percentage. This change in wire length introduces a change in reactance near the series-resonant points, and a change in resistance near parallel-resonant points. Both may be relatively large.

A better approximation occurs when the conductor length is kept constant. It is not easy to evaluate the precise error this causes, but it appears that it is no greater than the sum of the inherent error of MININEC plus the area error in the approximation.

It also appears that the inherent error will be around 5 to 10 percent if two simple rules are followed in setting up a MININEC analysis:

- use a minimum of two segments per section of conductor,
- use a minimum of four segments per halfwave of conductor.

Practically, MININEC gives extremely good accuracy. An antenna can easily depart from its ideal value by 20 percent or more because of neglected factors like supports, tower location, and feed-antenna interaction.

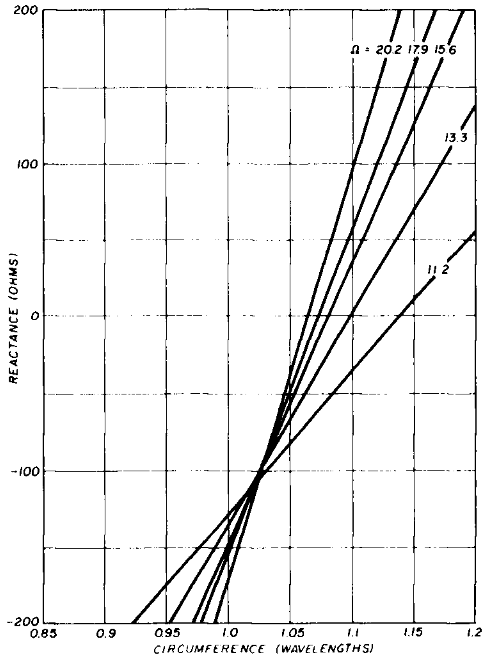


fig. 10. Feed or self-reactance of an octagon for circumferences around one wavelength. The point of equal values occurs at the same reactance as the circular loop but at a greater circumference. Curve slopes are almost identical.

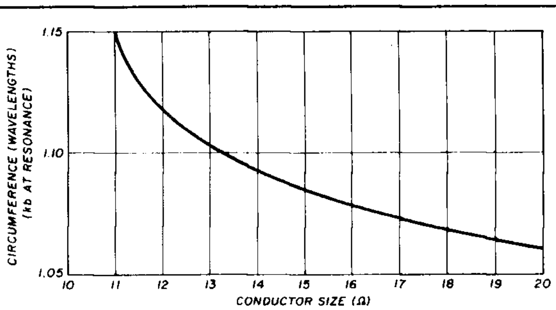


fig. 11. Resonant wavelength of an octagonal loop as a function of conductor thickness. The curve is similar to that of a circular loop, but differs by as much as ±6 percent.

octagonal loop

Calculated values for an octagonal loop should be a good approximation of circular-loop values. The octagon is also a useful antenna by itself. Probably the best-known type is the "Army Loop" — really a loop operating well below resonance, incorporating both a matching system and low-loss construction. **Figure 8** shows the basic factors involved, and allows visualization of the small area difference from the circular loop.

The properties of the octagon for conductor lengths close to 1 wavelength are summarized by **figs. 9A** and **9B** for drive resistance and reactance. These should be compared with **figs. 4A** and **4B** to see the validity of the octagon approximation to a circular loop. As expected from previous discussions, the agreement

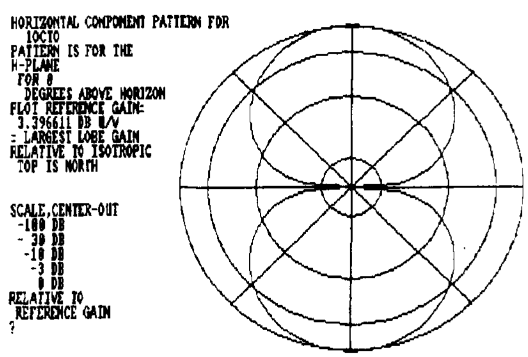


fig. 12. Horizontal radiation pattern for an octagon lying in a vertical plane and fed at the bottom. Only the horizontally polarized component is shown. The pattern is very similar to that of a dipole, but the gain is higher by about 1.25 dB.

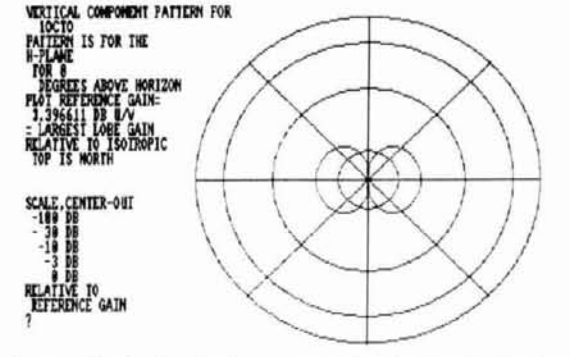


fig. 13. Vertical radiation pattern for the total radiation from an octagon in the plane of the loop. Maxima are along the line of maximum current. The minima represent the expected side radiation. (kb = 1.0, Ω = 17.9)

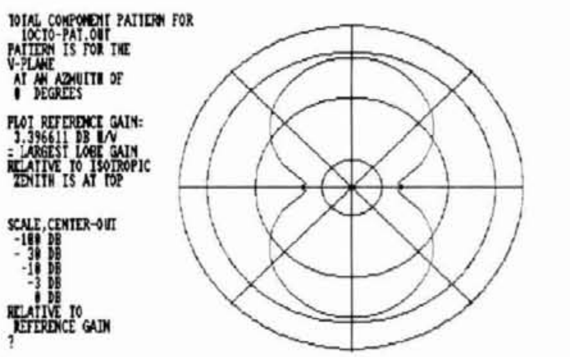


fig. 14. Horizontal radiation pattern for the vertically polarized component of an octagon lying in a vertical plane and fed at the bottom. This component is about 20 dB below the main lobe because of the currents on the vertical and inclined parts of the octagon.

for resistance is good — no more than a few percent difference. The agreement for reactance is not quite as good; this is the usual finding. However, neither error is beyond the expected range.

If you build a practical octagon antenna, use **figs. 10** and **11** if the antenna is to be resonant or nearly so. If it is away from resonance, the values from **table 1** give a good approximation. For best accuracy, the table values should be interpolated for the actual conductor diameter and equivalent radius. Alternatively, MININEC can be used to obtain values within 5 to 10 percent error.

patterns of circular/octagonal loops

Because the octagon has been found to be a good approximation to the circle, we can use its pattern as a close approximation to those of circular loops.

Figure 12 shows the MININEC calculated horizontal plane pattern for a 1-wavelength octagon fed at the center of the bottom segment — that is, with horizontal polarization. The lobes are very nearly the same as those of a dipole, but are slightly narrower to produce gain. The gain is 3.4 dB, essentially that of a circular Quad.

Figure 13 shows the vertical plane pattern in the plane of the loop, for the total radiation. **Figure 14** shows the horizontal plane pattern for vertical polarization. There is a major difference from dipole patterns in these two figures. The loop has an appreciable vertically polarized component, zero on axis and maximum at right angles to this. This component is not present with dipoles and its importance is not clear. It may have an adverse effect because of interference received on the vertically polarized side lobes (and back lobe in beams), or it may tend to reduce fading due to variation of incoming signals and path splitting. The component may also be a factor in the reputation of Quad loops as good performers at low height. More aspects of these questions will be discussed in other parts of this series.

Figure 15, provided for comparison, shows the calculated current on an octagon. Because the example is for a relatively thin conductor, there is close resemblance to the distribution for a transmission line.

Figure 16 compares the calculated gain of an octagon and a circular loop. Some of the variation is due to the difference in areas. The rest appears to be related to differences in the calculation method. Practically, the difference is negligible.

General conclusions from this comparison of circular and octagonal loops are that both have good performance, and that the data from one can be used for the other with little error.

We will use this last finding in part 3, which is devoted to arrays composed of circular and octagonal loops. Design data for arrays of two to twelve elements will be given.

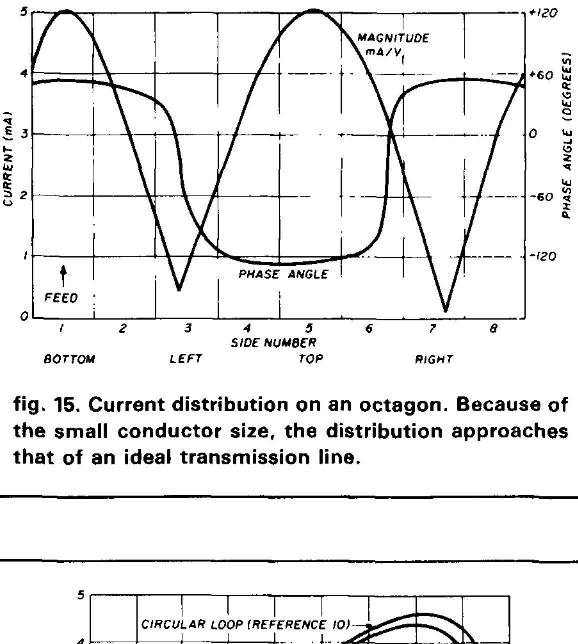


fig. 15. Current distribution on an octagon. Because of the small conductor size, the distribution approaches that of an ideal transmission line.

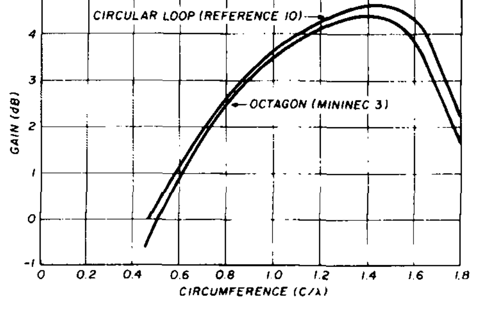


fig. 16. Calculated gain of a circular loop and an octagon. The difference is small — partly real and partly due to differences in the calculation technique. The curves do confirm the extra gain of a loop as compared with a dipole, and also indicate the added gain possible with off-resonance operation.

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ham radio

PART 3 — COMING UP IN OUR JULY ISSUE.

HAM RADIO TECHNIQUES

Bill Orr, W6SAI

a nifty bi-square beam for 10 or 12 meters

The miserable DX conditions at the bottom of the sunspot cycle are but a bad memory. True, the higher frequency bands tend to fizzle during the summer, but they'll be back again with a bang as soon as the cooler fall months roll around.

If you're interested in DX operation on either 10 or 12 meters, you'll eventually need a beam antenna. You can work a lot of "easy" DX with a dipole, but sooner or later you'll wish you had a beam for the more exotic DX stations. An easy solution is to buy a Yagi beam kit, but it's less expensive to build your own wire beam from scratch. Here's an inexpensive beam for your consideration.

The Bi-square beam (**fig. 1A**) is a derivation of the so-called "Lazy-H" array, a favorite of point-to-point stations in the maritime and fixed services. The Lazy-H consists of two half-wave dipoles in phase over a similar pair of dipoles. Spacing between the top and bottom dipole pairs is a half wavelength. Proper phasing of the pairs is achieved with a transposed open-wire transmission line fed at the center of the lower pair of dipoles with a quarter-wave, open-wire stub. The feedpoint impedance at the bottom of the stub is about 220 ohms.

A more practical version of the Lazy-H antenna is the Bi-square beam, shown in **fig. 1B**. This arrangement requires only a single center pole support. The Lazy-H dipole pairs are connected together at the outer tips, resulting in a diamond-shaped wire arrangement. You can eliminate the transposed line connecting the center of the pairs. The quarter-wave stub is retained.

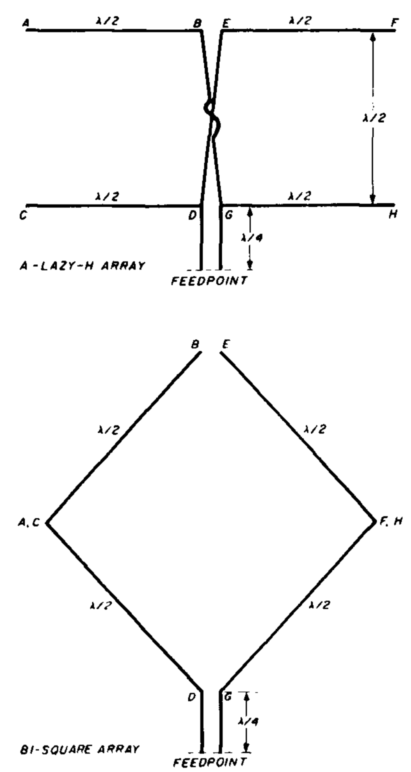


fig. 1 Simple bidirectional wire gain antennas: (A) lazy-H array and (B) bi-square array.

The feedpoint impedance at the bottom of the stub is close to 150 ohms. There is a reduction in feedpoint impedance because the top and bottom radiating elements of the Bi-square configuration are closer to each other than they are in the Lazy-H antenna.

The Bi-square radiation pattern is a figure eight (bidirectional) at right angles to the plane of the array. The power gain over a dipole located at the center height of the array is about 5 dB.

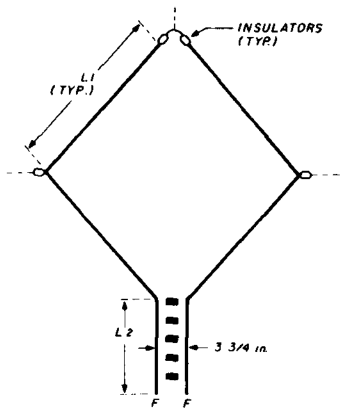
building the bi-square beam

The Bi-square is an easy, inexpensive beam to build. You'll need about 100 feet of No. 16 enamel or Formvar™ coated wire and four insulators. The quarter-wave stub needs five spreaders cut from 1/2-inch diameter phenolic (or plastic) rod. One of the spreaders serves as the bottom insulator for the antenna wires. The diamond-shaped antenna is open at the top (two insulators required). Overall height is a little less than 30 feet. I hung mine from a yard arm at the 45-foot level of my crank-up tower. The proximity of the metal tower to the plane of the loop didn't seem to cause any harm.

Dimensions for the 10- and 12-meter versions of the antenna are given in **fig. 2**. The sides are pulled out by ropes and tied off to convenient points on nearby trees. The bottom of the quarter-wave stub is about 7 feet above the ground.

The yard arm holds the loop about 3 feet away from the tower. The loop isn't quite in the vertical plane because I pulled the bottom of it 6 feet away from the tower in order to reach the bottom of the stub easily from the garage roof.

The Bi-square antenna's bandwidth is very broad; the antenna may be cut to the dimensions given without further ado. Purists may wish to trim the antenna to a specific frequency in the 10-meter band. Design frequencies for the antenna shown are 28.5 and 24.95 MHz. The 10-meter antenna covers the whole band with an SWR of less than 1.5:1 — quite an achievement!



FREQ. (MHz)	L1	L2
28.5	16 ft. 9 in.	8 ft. 4 in.
24.95	19 ft. 2 in.	9 ft. 5 in.

fig. 2. Bi-square antenna dimensions for 10- and 12-meter bands feedpoint impedance is about 150 ohms.

adjusting the antenna to frequency

It's easy to set the resonant frequency of the antenna "on the nose." The bottom of the stub (F-F) is shorted by a jumper that has a one-turn loop in the center. The loop is just big enough to fit over the coil of a dip oscillator. My shorting bar is made of two interconnected copper alligator clips so I can move it up and down the stub for adjustment. The dip oscillator is monitored in a nearby receiver. Move the shorting bar up and down the stub, an inch or so at a time, until the resonant frequency falls where you want it. Finally, cut the stub to the determined length.

matching antenna to feedline

As I stated earlier, the feedpoint impedance of the antenna is about 150 ohms. The antenna is symmetrical with respect to ground, and the feedpoint is balanced to ground. Two transformations are required to match the antenna to a 50-ohm unbalanced (coaxial) line. The 50-ohm point is first transformed from unbalanced to balanced by a 1:1 balun. The 50-ohm balanced condition is then transformed to 150 ohms. The first transformation is easy; I use a "Bencher ZA-1A" air-core balun which provides an excellent balance in the 10-meter region.

The transformation from 50 ohms to 150 ohms can be done in a number of ways. One is to use a ferrite toroid transformer (**fig. 3**). Take a core 2.4 inches in outer diameter and 0.5 inch high (Amidon FT-240-67, or equivalent) with a permeability of 40. Sand it to remove rough edges, and then wrap it with a layer of electrical vinyl tape. Wind 18 turns of No. 14 enamel wire around the core, tapped four turns from each end. Space the winding around the entire core. Fasten the completed transformer to a phenolic mounting plate with epoxy cement, and mount the assembly in a waterproof box for protection from the weather.

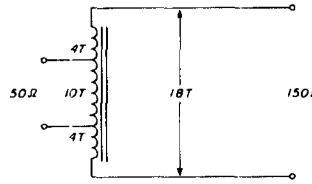


fig. 3. A 1:3 balanced transformer wound on ferrite toroid (Amidon FT-240-67).

a linear matching transformer

The second matching scheme uses a linear transformer, (**fig. 4**). The design is based on a balanced L-network. The circuit (**fig. 4A**) was built using a receiving-type variable capacitor for initial tests. The dimensions shown allow adjustment of the capacitor which quickly drops the SWR on the transmission line to unity at the design frequency of the antenna. The last step is to replace the variable capacitor with a fixed one and substitute a section of transmission line for the network inductors (**fig. 4B**). This works like a charm. A 50-μF, 5-kV ceramic capacitor (Centralab 850S-50Z, or equivalent) is substituted for the variable unit. Place it in a plastic refrigerator jar to keep moisture away. The short line section is made up in the same manner as the quarter-wave stub.

results

For a few days the dipole was left in position as a comparison with the Bi-square. In all tests, the Bi-square outperformed the dipole (usually between one and two S-units on transmit). On receive, signals that were almost in the noise were perfectly readable on the Bi-square antenna. No doubt about it, the Bi-square delivers the goods!

a 15-meter version?

The Bi-square should work well on 15 meters if you have the space. Multiply all 10-meter linear dimensions by 1.34 to get antenna size for this band.

W5LDA 160-meter antenna

Jim, W5LDA, has an interesting 160-meter antenna that incorporates a simplified feed system (**fig. 5**). He uses

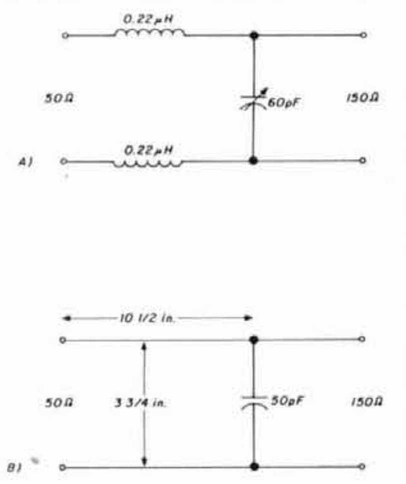


fig. 4. Linear matching transformers. (A) L-network uses lumped L and variable C (five turns, 1/2-inch I.D., No. 14 wire, eight turns per inch). (B) Linear network transformer made of No. 16 wire. Ends of wire are bent in 1 inch to connect to balun.

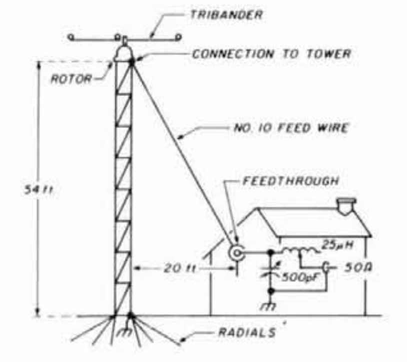


fig. 5. 160-meter folded unipole antenna at W5LDA. Rotor cable and coax for beam is strapped to tower leg.

his 54-foot tower (with a triband Yagi atop it) as a vertical, top-loaded radiator. Rather than fooling around with a gamma match on the tower (which can prove to be very tricky), Jim made his tower into a voltage-fed unipole antenna. He fastened a wire to the top, brought it off at an angle, and voltage fed the bottom end. The natural resonance of the top-loaded tower is such that only a simple matching network is required.

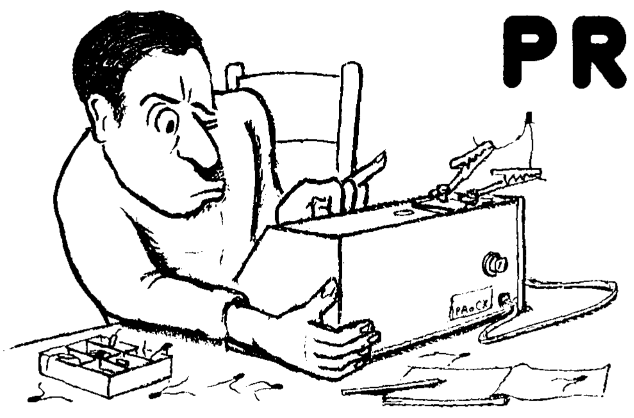
The base of the tower, as well as the shield of the coax running to the beam, are grounded at the tower base. Each lead of the rotor cable (not shown) is bypassed to ground at the tower base with a 0.01-μF, 1.6-kV disc capacitor. The leads are also bypassed to the tower at the rotor. (Jim learned the hard way that bypassing is important, after he burned out the rotor potentiometer atop the tower running 1500 watts on 160 meters!)

The coax and rotor cables are buried in a hose and run to the shack. Twenty radials, each 65 feet long, are fanned out on the surface of the ground beneath the tower.

The end of the wire is at a high voltage point and is brought into the station via a ceramic feedthrough insulator. A simple L-network matches the antenna to 50-ohm coax running to the operating position.

The antenna is very high-Q (narrow bandwidth); the network must be readjusted for a frequency change. It is possible to achieve 80-meter operation of the antenna by retuning the network.

ham radio



PRACTICALLY SPEAKING

Joe Carr, K4IPV

"ferriting" out the problem

Ferrite refers to materials that behave similarly to powdered iron compounds. They are used in radio equipment as inductors and transformers. Although they were made originally from powdered iron (and indeed the name *"ferrite"* still implies iron), many modern materials are made of other compounds. According to the Amidon Associates, ferrites with a permeability of 800 to 5000 have manganese-zinc composition, while cores with permeabilities of 20 to 800 are of nickel-zinc.[1] The latter are useful in the 0.5- to 100-MHz frequency range of interest to most Amateurs.

toroidal cores

This month's column will answer your questions about ferrite cores, winding toroidal cores, and using ferrite inductor and transformer cores.

A toroid is a "doughnut" shaped object, so a toroidal core is an inductor or transformer form made of a ferrite material in the shape of a doughnut. Core nomenclature provides useful information about shape, size, and type of material. For example the number *FT—xx—nn* means a ferrite toroid (FT) with an "xx" size, and an "nn" material type. The "F" in "FT" is sometimes deleted on parts lists, and the core defined as a "T-xx-nn."

Amidon has a chart that provides dimensions, a description of the properties of the different types of material, and other physical data. Some of these data are also found in *The 1988 ARRL Handbook for the Radio Amateur*[2], beginning on page 2-32 (the same material appeared in earlier editions as well).

Table 1. Standard Toroid Core Sizes ("xx")

Core size	OD (inches)	ID (inches)	Thickness (inches)
23	0.230	0.120	0.060
37	0.375	0.187	0.125
50	0.500	0.281	0.188
50A	0.500	0.312	0.250
50B	0.500	0.312	0.500
82	0.825	0.520	0.250
87A	0.870	0.540	0.500
114	1.142	0.750	0.295
114A	1.142	0.750	0.545
130	1.300	0.780	0.437
150	1.500	0.750	0.250
150A	1.500	0.750	0.500
193	1.930	1.250	0.750
200	2.000	1.250	0.550
240	2.400	1.400	0.500

Table 2. Properties of Core Types

Material type	Color code	μ	Frequency (MHz)
41	Green	75	—
3	Grey	35	0.05—0.5
15	Red/ White	25	0.1—2
1	Blue	20	0.5—5
2	Red	10	1—30
6	Yellow	8	10—90
10	Black	6	60—150
12	Green/ White	3	100—200
0	Tan	1	150—300

Tables 1 and **2** are derived in part from both Amidon and ARRL sources; **table 1** shows the sizes and **table 2** the properties of various popular toroids. These tables do not contain an exhaustive list of the variety of toroids available or all the properties of the toroids mentioned. Using the nomenclature mentioned above and the tables, you can see that a T-50-2 core (which might be called for in the parts list of a *ham radio* article) refers to a core that is useful from 1 to 30 MHz. It has a permeability of 10, is painted red, and has the following dimensions: OD = 0.500 inches, ID = 0.281 inches, and height (i.e. thickness) = 0.188 inches.

toroidal transformers

One reader asked me about the winding protocol for toroidal transformers seen in Amateur books and magazine articles. My correspondent included a partial circuit (**fig. 1A**) as an example of his dilemma. He wanted to know how to wind it and proposed a couple of methods. At first I thought the answer was obvious, then I realized that I was wrong — to many people it is not.

All windings are wound together in a "multifilar" manner. If there are three, we are talking about "trifilar" windings. **Figure 1B** shows the trifilar winding method. For clarity's sake, I have shown all three wires differently. Because most of my projects use No. 26, 28, or 30 enameled wire to wind coils, I keep three colors of each size on hand and use a different color for each winding.

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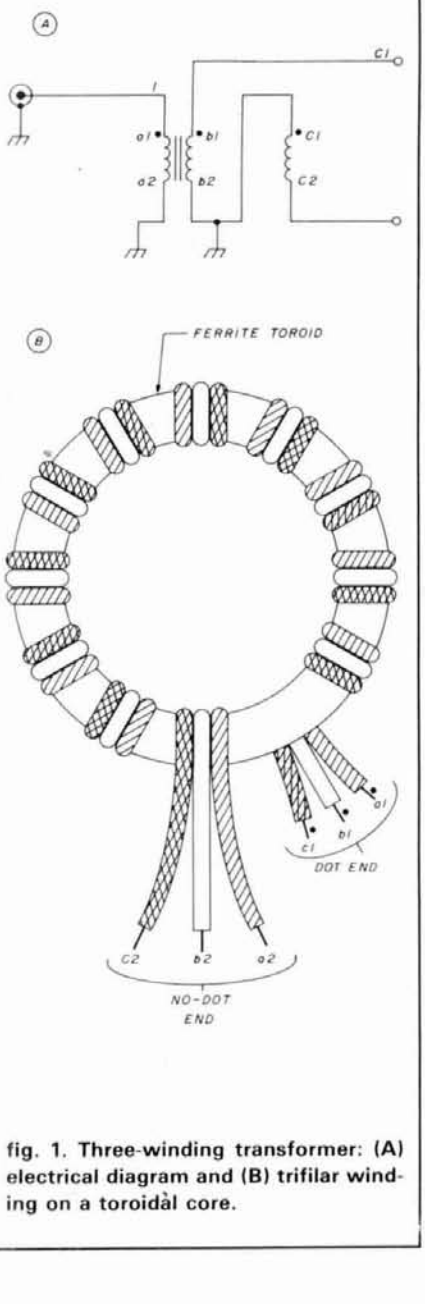

fig. 1. Three-winding transformer: (A) electrical diagram and (B) trifilar winding on a toroidal core.

The dots in the schematic and on the picture identify one end of the coil windings. The "dot" and "no-dot" ends are different from each other, and it usually makes a difference to circuit operation (signal phasing) which way the ends are connected into the circuit.

Figure 2 shows two accepted methods for winding a multifilar coil on a toroidal core. **Figure 2A** is the same method as in **fig. 1B**, but shows an actual toriod rather than a pictorial representation. As previously shown, the wires are laid down parallel to each other. The method in **fig. 2B** uses twisted wires. The three wires are chucked up in a drill and twisted together before being wound on the core. With one end of the wires secured in the drill chuck, anchor the other end to something that will hold it taut. (I use a bench vise.) Turn the drill on slow speed and let the wires twist together until you achieve the desired pitch.

Be very careful when performing this operation. If you don't have a variable speed electric drill that runs at very slow speed, use an old-fashioned manual drill. Remember to wear eye protection if you use an electric drill. If the wire breaks, or gets loose from its mooring at the end opposite the drill, it will whip around wildly until the drill stops. That whipping wire will cause painful welts on the skin, and can certainly cause eye damage.

Of the two methods for winding toroids, the method shown in **figs. 1B** and **2A** is preferred. When winding toroids, at least those of relatively few windings, pass the wire through the "doughnut" hole until the toroid is close to the midpoint of the wire. Loop the wire over the outside surface of the toroid, and pass it through the hole again. Repeat this process until the correct number of turns is wound onto the core. Be sure to press the wire against the toroid form and keep it taut as you wind the coils.

Enameled wire is usually used for toroid transformers and inductors, and this can lead to problems. The enamel can chip causing the copper conductor to contact the core. On larger cores, like those used for matching transformers and baluns at kilowatt power levels, the practical solution is to wrap the bare toroid core with a layer of fiber glass packing tape. Wrap the tape exactly as if it were wire, but overlap the turns slightly to ensure covering the entire surface of the core.

On some projects, particularly those in which the coils and transformers use very fine wire (like No. 30), I have found that the wire windings tend to unravel after the process is complet-

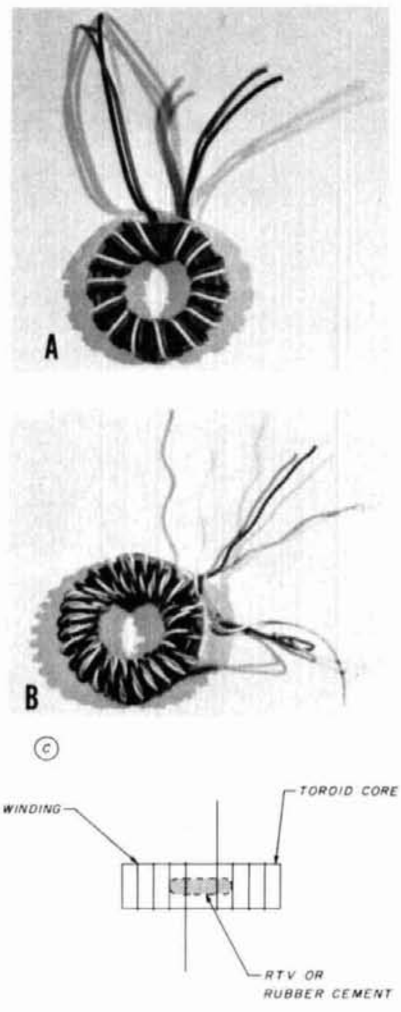

fig. 2. Winding methods and hints: (A) conventional turns, (B) twisted leads, and (C) secured winding with dab of glue or sealer.

ed. To prevent this, place a tiny dab of rubber cement or silicone sealer at the ends of the windings (see **fig. 2C**).

mounting toroid cores

Now that you have a properly wound toroidal inductor or transformer, it is time to mount it in the circuit. There are three easy ways to do this. If the wire is strong enough, use the wire connections to the circuit board or terminal strip to support the component. If this is not satisfactory (and in mobile equipment or wherever else vibration is a factor it won't be), try laying the toroid flat on the board and cementing it in place with silicone

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seal or rubber cement. For the third method, drill a hole in the wiring board and use a screw and nut to secure the toroid. *Do not use metallic hardware for mounting the toroid!* Metallic fasteners will alter the inductance of the component and possibly render it unusable. Use nylon hardware for mounting the inductor or transformer.

Table 3. Common A_L values.

Core	Core Materials Type (Mix)								
Size	26	3	15	1	2	6	10	12	0
12	—	60	50	48	20	17	12	7	3
16	—	61	55	44	22	19	13	8	3
20	—	90	65	52	27	22	16	10	3.5
37	275	120	90	80	40	30	25	15	4.9
50	320	175	135	100	49	40	31	18	6.4
68	420	195	180	115	57	47	32	21	7.5
94	590	248	200	160	84	70	58	32	10.6
130	785	350	250	200	110	96	—	—	15
200	895	425	—	250	120	100	—	—	—

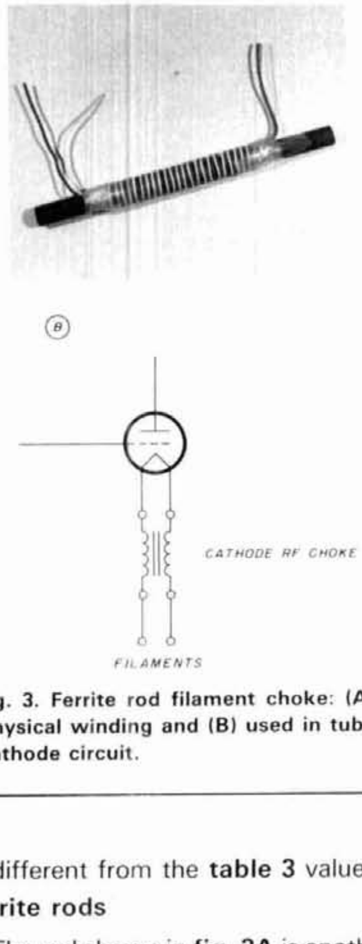

fig. 3. Ferrite rod filament choke: (A) physical winding and (B) used in tube cathode circuit.

how many turns to use?

Three factors must be considered when making toroid transformers or inductors: toroid size, core material, and number of turns of wire. The toroid size is selected as a function of power handling capability or convenience. The core material is selected according to the frequency range of the circuit. The only thing left to vary is the number of turns. The size and core material yield a figure called the A_L factor. These values are given for several popular toroidal cores in **table 3**. The required value of inductance and the A_L factor are related by the following equation:

$$N = 100\sqrt{L/A_L} \qquad (1)$$

Where:

N is the number of turns

L is the inductance in microhenries

A_L is the core factor in microhenries per 100 turns

EXAMPLE

Calculate the number of turns required to make a 5-μH inductor on a T-50-6 core. From **table 3** we see that the A_L factor is 40.

Solution:

$$N = 100\sqrt{L/A_L} \qquad (2)$$
$$= 100\sqrt{5/40} = 35$$

Don't take the equation value too seriously; a wide tolerance exists on Amateur-grade ferrite cores. While this isn't too much of a problem when buildng transformers, it can be critical when making inductors for a tuned circuit. If you find that the tuned circuit takes considerably more or less capacitance when called for in the standard equation, and all of the stray capacitance is properly taken into consideration, then it may be that the actual A_L value of your particular core is different from the **table 3** value.

ferrite rods

The rod shown in **fig. 3A** is another form of ferrite core available on the Amateur market. This type of core is used to make rf chokes, like the one used in the vacuum tube filament lines of a linear amplifier power tube (**fig. 3B**). The two windings are made in a bifilar manner over the ferrite rod. The wires used are heavy enough to carry the filament current of the tube. As in the toroidal transformer, I used two different wire colors to discriminate between windings.

Ferrite rods are also used in receiving antennas. Although few Amateurs have them, there are places where a ferrite rod antenna (or "loopstick") is used. For example, ferrite loops are common in radio direction-finding antennas. Some Amateurs report that they use a loopstick receiving antenna when operating on crowded bands like 40 or 75 meters. The small loopstick is extremely directional and is capable of nulling out interfering signals. Of course, one would not want to use the loopstick for transmitting, so there must be some means for switching between transmit and receive functions.

mounting ferrite rods

Ferrite rods can be mounted several ways; two of them are analogous to the methods used on toroids. You can mount the rod using either its own wires for support or a dab of cement or silicone sealer to fasten it to the board. Although you can't use simple nylon screws the way you can on toroids, you can use insulating cable clamps to secure the ends of the rod to the board.

Questions, suggestions, and criticisms are welcomed. Send them to: Joe Carr, K4IPV, POB 1099, Falls Church, Virginia 22041.

references

1. Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607

2. *The 1988 ARRL Handbook for the Radio Amateur*; $20.95 plus $3.50 shipping and handling from *ham radio* Bookstore, Greenville, New Hampshire 03048.

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Yagi vs. Quad, Part 2

Optimized forward gain comparison

In part 1, we developed a means of analyzing quad antennas based on a mutual impedance versus spacing relationship and a fixed single quad element pattern. In part 2, we will examine different quad element configurations on a variety of boom lengths and attempt to maximize forward gain through fine tuning of the element lengths. Boom length will be limited to 1 wavelength or less because this is about the longest practical boom for 15, 20, or 40 meters.

The intent here is to answer the questions: What is the best possible forward gain we can squeeze out of a quad of a given boom length, and how much has the optimized gain improved from where we started? Note that we are not taking front/back discrimination or bandwidth into consideration. Both parameters are important, but both detract from maximized forward gain. Finally, we can compare the computed maximized gain for a quad against the maximized forward gain for a Yagi.

computational methodology

Antennas are modeled in free space using the assumptions outlined in part 1*. The mutual impedance between elements is assumed to be independent of element length. The self-impedance is approximated from interpolation of measured values. Gain is calculated from integration of field pattern over a sphere, and comparison of the forward field against the average power. Field points are updated every 10 degrees (both phi and theta, spherical coordinates).

Initial antenna design is based on general *ARRL Antenna Handbook* principles. Reflector length is 1030/f = 1.05 wavelengths. Driven elements are 1005/f = 1.021 wavelength. Directors are all 975/f = 0.991 wavelengths, and all elements are spaced equally along the boom. Starting with the reflector, the element length will be increased by 0.0025 wavelength and the gain again calculated. If the forward gain improves by at least 0.005 dBi, that element is incremented again in the same direction; if not, the element will be shortened until the gain starts to fall off. Once the reflector is optimized for forward gain, the same procedure is applied to the directors in order, and the process is repeated until no significant gain increase occurs with element change. The elements, except for the ones at the end of the boom, will then be moved along the boom in 0.0025 wavelength increments to attempt a further gain increase.

two-element quad results

The optimized gain was about 8 dBi on a 0.115 or 0.172 wavelength boom (8 and 12 feet on 20 meters). Adding a third element was worthwhile, as gain increased to about 10 dBi for boom lengths of 0.258, 0.30, 0.343, 0.430, and 0.516. Initial performance of larger antennas was better. Quads on a 0.688 wavelength boom (48 feet on 20 meters) showed an initial gain of about 11 dBi which increased to just under 12 dBi with tuning. Quads on a 0.86 wavelength boom (60 feet on 20 meters) could be tuned for over 12 dBi, and on a 1.031 wavelength boom (72 feet on 20 meters) peaked at 12.7 dBi.

The relationship between maximum gain and boom length is compared against similar data developed for Yagi antennas[1] (**fig. 1**). The two-element quad is slightly better than a two-element Yagi (by 0.5 dBi or so) but worse than a three-element full-size Yagi (by 1.5 dBi), and a three-element quad is slightly better than a three- or four-element Yagi (again by 0.5 dBi). The quad antennas did not show the same staircase gain phenomenon observed with Yagi antennas[1], and *forward gain rose smoothly with increases in boom length out to 1 wavelength.* With Yagi antennas the gain versus boom length relationship is not smooth, but increases rapidly over a small extension in boom length and then plateaus. Thus, Yagi and quad antennas performed similarly out to a 24-foot boom on 20 meters, but past this point quad forward gain continued to increase and Yagi forward gain plateaued (**fig. 2**). Between 24- and 48-foot booms the quad showed up to 2.5 dB gain over the same size Yagi. Above 48

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*Yagi vs Quad, Part 1, *ham radio*, May 1988, pg. 68.

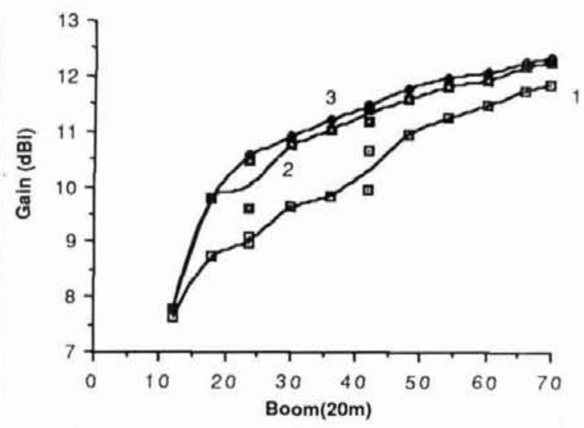


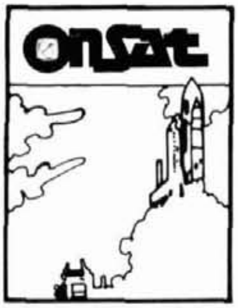
fig. 1. Forward gain versus boom length in feet for different 20-meter antennas. *Line 1 (open squares) shows the* original antennas; line 2 (filled squares) gives results after forward gain optimization by perturbation of element lengths; line 3 (filled triangles) shows what happens after *element and spacing perturbation.* Note the gain improvement with optimization was about 1 dB.

feet the Yagi closed the gap, reducing the quad advantage to about 1.5 dB.

Several conclusions are suggested by these results: First, quad antenna gain, like Yagi gain, is basically a function of boom length and not the number of elements, as long as "enough" elements are used. Second, the quad may have 2 to 2.5 dB greater gain than a well-tuned Yagi, but this occurs only at certain boom lengths. The quad should work about 0.5 to 1.5 dB better for the same boom length. Third, the average gain increase expected with fine tuning is about 0.45 dB (range 0.0-1.55 dBi).

Lindsay's 440-MHz experimental results initially suggested that quads were somewhat better than Yagi antennas on the same-sized boom[2]. However, experimental uncertainties led to questions about his conclusions. For instance, it was difficult to measure gain accurately on the basis of input power because of inefficiencies in the coupling system.[3] This may have influenced Lindsay's measurements from the start, since he measured the gain of a quad loop as being 2 dB better than a dipole. The actual value, according to several sources, should be 1 dB. Another variable in Lindsay's study involved the lengths of Yagi and quad elements; they were cut by formula and not optimized for any particular factor like forward gain. Because fine tuning of a Yagi or quad may add another 1 dB or so to the forward gain, any gain difference between Yagi and quad antennas is reduced to the experimental noise.

Not all people have found quad antennas to be as good as Yagis. Driving around California with a 70-foot portable reference antenna, Wayne Overbeck com-



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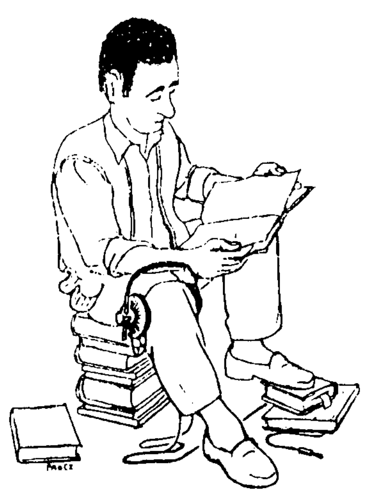

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Table 1. Element length, spacing, forward gain and front to back (at the horizon) for two, three, and four elements on 0.115, 0.258, 0.343, 0.430 and 0.516 wavelength booms (8, 12, 18, 20, 24, 30, and 36 feet on 20 meters). Row S is the starting antenna and row O is after forward gain optimization. The element length is specified first and the element position is in parentheses. R is the reflector; De is the driven element; D1..Dn are directors. Gain is specified over isotropic.

Element Length (boom position) (R)	(De)	(D1)	(D2)	(D3)	(D4)	(D5)	(D6)	Gain(dBi)	F/B(dB)
S 1.050(0)	1.021(.12)							7.64	20.8
O 1.035(0)	1.021(.12)							8.00	12.0
S 1.050(0)	1.021(.17)							7.65	14.4
O 1.037(0)	1.021(.17)							7.78	9.7
S 1.050(0)	1.021(.13)	0.991(.26)						8.76	13.0
O 1.030(0)	1.021(.13)	1.013(.26)						9.84	10.2
S *1.050(0)*	*1.021(.15)*	*0.991(.30)*						8.99	15.2
O 1.035(0)	1.021(.15)	1.013(.30)						9.72	9.1
S 1.050(0)	1.021(.12)	0.991(.23)	0.991(.34)					9.03	10.0
O 1.030(0)	1.021(.12)	1.008(.23)	1.021(.34)					10.42	14.0
S 1.050(0)	1.021(.11)	0.991(.22)	0.991(.33)	0.991(.43)				9.65	15.0
O 1.027(0)	1.021(.11)	1.011(22)	1.021(.33)	1.016(.43)				11.22	16.4
S 1.050(0)	1.021(.13)	0.991(.26)	0.991(.39)	0.991(.52)				10.28	25.5
O 1.037(0)	1.021(.13)	1.011(.26)	1.021(.39)	1.021(.52)				11.52	17.0
S 1.050(0)	1.021(.17)	0.991(.34)	0.991(.52)	0.991(.69)				10.97	25.4
O 1.039(0)	1.021(.15)	1.011(.30)	1.021(.52)	1.003(.69)				11.78	16.0
S 1.050(0)	1.021(.17)	0.991(.34)	0.991(.52)	0.991(.69)	0.991(.86)			11.48	12.6
O 1.040(0)	1.021(.17)	1.001(.34)	0.963(.52)	1.016(.69)	1.011(.86)			11.99	13.6
S 1.050(0)	1.021(.15)	0.991(.29)	0.991(.44)	0.991(.59)	0.991(.74)	0.991(.88)	0.991(1.03)	11.96	43.8
O *1.035(0)*	*1.021(.15)*	*1.006(.29)*	*0.971(.44)*	*0.988(.59)*	*0.991(.74)*	*0.988(.88)*	*1.026(1.03)*	12.59	16.8

pared low-angle signal strengths of Yagi and quad antennas at various QTHs against a portable reference antenna at the same height as the test antenna[3]. A review of his results showed that quads were the same as or inferior to Yagi antennas, and certainly not 2 dB better than similar Yagi antennas. Why didn't the quads perform better?

Theoretical results presented here suggest that quad antennas with correct coupling should be at least as good as Yagis, and certainly not worse. But two complicating factors were not included in the computer model: the efficiency of the feed system and the effects of nonresonant elements in the vicinity of the antenna. Most of the antennas showing the best performance in Overbeck's study were monoband Yagis with a double driven element. This feed system, popularized by KLM, is known for its wide bandwidth and low SWR, and is also said to give excellent results on quad antennas.[5] Perhaps part of the answer lies in a lower feed efficiency for direct feed quads. Attaching the coax to the quad loop is the most common method of feeding quads. Most quad antennas are implemented as tribanders with wires for other bands in close proximity to the desired antenna. It may be that these other wires significantly degrade performance. Both of these factors deserve a closer look.

I didn't intend to provide optimum dimensions for quad loops, and a caveat is in order if you wish to use the dimensions provided in Table 1. Although I believe in the results of computer modeling, the dimensions should be trusted only after confirmation on an antenna test range. I have not done this, and the actual performance peaks may be different. The constructed quad loops should have the same reactance as the computer antenna. The best way to do this is to direct-

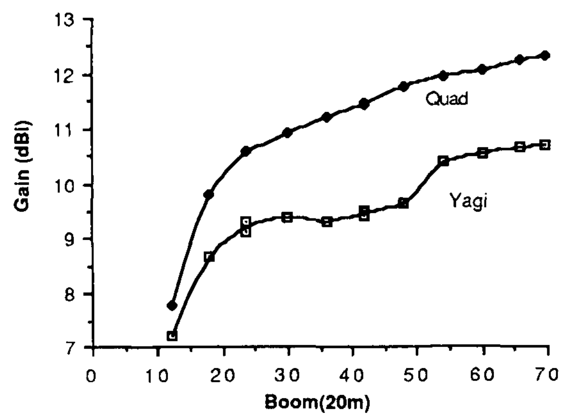

fig. 2. Forward gain versus boom length in feet for gain-optimized 20-meter Yagi and quad antennas. Open squares represent gain-optimized Yagi antennas from a previous article.[1] The quad line shows optimized gain from *fig. 1*. Quad antennas may have a 2-dB advantage over a Yagi on the same length boom, but only at specific boom lengths.

ly measure reactance with an impedance bridge and trim the element accordingly.

summary

Quad antennas offer a theoretical 0.5 to 2.5 dB advantage over a Yagi of the same boom length. Like the Yagi, a gain increase of about 1 dB may be obtained through fine tuning of the elements. The increase in forward gain as one goes to longer boom lengths is smoother for quad antennas than for Yagis, and quads do not show the gain plateau seen with Yagi antennas. Finally, the quad should be significantly better than the Yagi, especially between boom lengths of 25 and 45 feet on 20 meters. However, unknown variables such as feed system efficiency or the effect of other wires in close proximity to the quad may detract from its theoretical performance.

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4. B. Myers, "The W2PV Four-element Yagi," *QST*, October 1986, page 15-19.
5. R. Martinez, "The Evolution of the Four-element Double-driven Quad Antenna," *CQ*, December 1983, page 30-36.
6. J.L. Lawson, "Yagi Antenna Design: Quads and Quagis," *ham radio*, September 1980, page 37-45.

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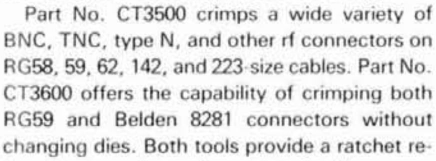

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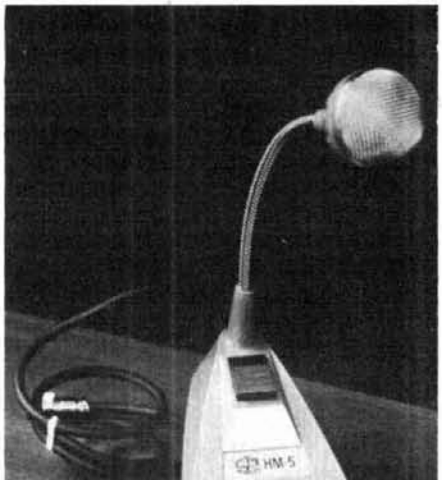

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Macket's suggested retail price is $39.95. The program, developed by 'S Fine Software' is available from Pac-Comm Packet Radio Systems, Inc., 3652 W. Cypress St., Tampa, Florida 33607.

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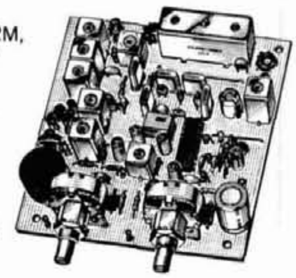
VHF OR UHF

FEATURES:
- **SENSITIVITY SECOND TO NONE! GaAsFET front end** on vhf models gives 12dB SINAD of 0.12uV (vhf), 0.15uV (220). UHF model 0.25uV std, 0.1uV with optional helical resonator preamp.
- **SELECTIVITY THAT CAN'T BE BEAT!** Both 8-pole xtal filter & ceramic filter for **> 100dB at only ±12kHz.** Helical resonator front end to combat desense & intermod.
- **CLEAN, STABLE TRANSMITTER,** up to 18W output standard; 50W with accessory power amplifier.
- **FCC TYPE ACCEPTED** for commercial high band and uhf.
- **Courtesy beep,** field-programmable CWID, flutter-proof squelch, automatic frequency control to compensate for off-frequency transmitters (all standard features).
- **Full range of options** available, such as autopatch, phone line or radio remote control, sub-audible tones, duplexers.

HIGH PERFORMANCE TRANSMITTERS & RECEIVERS FOR REPEATERS AUDIO & DIGITAL LINKS, TELEMETRY, ETC.

- **FM EXCITERS:** Kits $99, W/t $179. 2W continuous duty. TCXO & xtal oven options available.
- **TA51 for 10M, 6M, 2M, 150-174, 220 MHz.**
- **TA451 for uhf.** FCC type accepted for commercial bands.
- Call for latest information on 900 MHz transmitters.
- **VHF & UHF AMPLIFIERS.** For FM, SSB, ATV. Output from 10 to 50 Watts. Several models, kits starting at $79.



- **R144/R220 FM RECEIVERS** for 2M, 150-174, or 220 MHz. **GaAs FET** front end, 0.12uV sensitivity! Both crystal & ceramic filters plus helical resonator front end for exceptional selectivity: > 100dB at ±12kHz (best available anywhere)! Flutter-proof squelch. AFC tracks *drifting transmitters.* Kit $149, w/t $229.
- **R451 UHF FM RCVR.** Similar to above. Tuned line front end, 0.25uV sens. (0.1uV with optional hel. res. preamp). Kit $149, w/t $229.
- **R901 FM RCVR FOR 900 MHZ.** Triple-conversion, GaAs FET front end, 0.2uV sens. Kit $169, w/t $259.
- **R76 ECONOMY VHF FM RCVR** for 10M, 6M, 2M, 220. Without hel res or afc. Kits only $129.
- **Weather satellite & AM Aircraft receivers also avail.**

FCC TYPE-ACCEPTED TRANSMITTERS & RECEIVERS AVAILABLE FOR HIGH-BAND AND UHF. CALL FOR DETAILS.

GaAs FET PREAMPS
at a fraction of the cost of comparable units!

LNG -(*)
GaAs FET PREAMP

ONLY $59! Wired/tested

FEATURES:
- **Very Low Noise:** 0.7dB VHF, 0.8dB UHF
- **High Gain:** 13-20dB, depending on frequency
- **Wide Dynamic Range:** to resist overload
- **Stable:** new-type dual-gate GaAs FET

* *Specify tuning range desired: 26-30, 46-56, 137-150, 150-172, 210-230, 400-470, or 800-960 MHz.*

LNW -(*)
MINIATURE GaAs FET PREAMP

ONLY $24/kit, $39 Wired/tested

GaAs FET Preamp similar to LNG, except designed for **low cost & small size.** Only 5/8"W x 1-5/8"L x 3/4"H. Easily mounts in many radios.

* *Specify tuning range desired: 25-35, 35-55, 55-90, 90-120, 120-150, 150-200, 200-270, or 400-500 MHz.*

LNS-(*)
IN-LINE PREAMP

ONLY $79/kit, $99 Wired/tested

GaAs FET Preamp with features similar to LNG series, except **automatically switches out of line during transmit.** Use with base or mobile transceivers up to 25W.

**Specify tuning range desired: 120-175, 200-240, or 400-500 MHz.*

HELICAL RESONATOR PREAMPS

Low-noise preamps with helical resonators **reduce intermod & cross-band** interference in critical applications.

MODEL HRA-(*), $49 vhf, $84 uhf.

* *Specify tuning range desired: 143-150, 150-158, 158-162, 162-174, 213-233, 420-450, 450-465, or 465-475 MHz.*

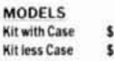
NEW HIGH-SPEED DIGITAL RF LINKS

You've waited a long time for a simple, reliable, **low-cost 9600 baud PACKET NETWORKING** *system.* Now you've got it! Our new MO-96 MODEM and direct FSK Transmitters and Receivers for 220 or 440 MHz interface directly with most TNC's. Fast diode switched PA's output 15 or 50W. **Call for complete info on the right system for your application.**

ACCESSORIES

- **COR-3** Kit. Control ckts and audio mixers needed to make a repeater. Tail & time-out timers, local spkr ampl, courtesy beep $49
- **CWID** Kit. Field programmable, timers, the works $59
- **TD-2 DTMF DECODER/CONTROLLER** Kit. Full 16 digits, switches 5 functions, toll call restrictor, programmable, much more. Great for selective calling too! $79
- **AP-3 AUTOPATCH** Kit. Use with above for repeater autopatch. Reverse patch and phone line remote control std. ... $79
- **AP-2 SIMPLEX AUTOPATCH TIMING BOARD** Kit. *Use with* above for simplex autopatch $39
- **MO-202 FSK DATA MODULATOR** Kit. Run up to 1200 baud digital signals *through any fm transmitter with* full handshakes. Radio link computers, telemetry gear, etc. $39
- **DE-202 FSK DATA DEMODULATOR** Kit for rcvr end of link $39

RECEIVING CONVERTERS



		Antenna Input Range	Receiver Output
VHF MODELS		28-32	144-148
		50-52	28-30
		50-54	144-148
		136-138	28-30
		144-146	28-30
		145-147	28-30
Kit with Case	$59	146-148	28-30
Kit less Case	$39	220-222	28-30
		220-224	50-54
Wired w/case	$89	222-224	28-30
UHF MODELS		432-434	28-30
		435-437	28-30
Kit with Case	$69	432-436	144-148
Kit less Case	$49	432-436	50-54
Wired w/case	$99	439.25	61.25
		902-928	422-448
		902-922	430-450

See catalog for full line of 2w transmitting converters for vhf & uhf. Kits only $79. Linear Amplifiers avail. up to 50 w.

- **Send $1 for 36 page catalog by return mail.** (Send $2.00 or 4 IRC's for overseas mailing)
- **Order by phone or mail • Min $3 S & H per order**
- **Use Visa, Mastercard, Check, or UPS COD.**

Our 25th Anniversary
hamtronics, inc.
65-H MOUL ROAD•HILTON NY 14468-9535
Phone: 716-392-9430 Hamtronics® is a registered trademark

NEW products

The new 5D-FB 50-ohm coax, usable above 1200 MHz, is a solid copper center conductor with a PEF insulation, foil wrapped, TFE insulation, full braid, and black PVC weather jacket. N type connectors are available.

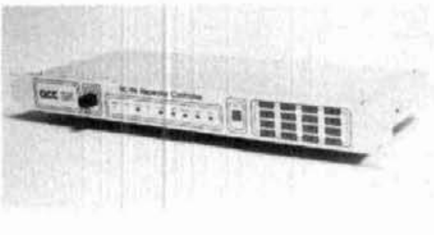
For information on the Tri-Band antennas, Tri-Plexers, and 5D-FB coax contact NCG Company, 1275 N. Grove Street, Anaheim, California 92806.

Circle #305 on Reader Service Card.

RC-96 repeater controller

Advanced Computer Controls, Inc. offers the RC-96 Repeater Controller. Remote programming lets the operator make changes to his repeater easily without a trip to the site.

The '96 features autopatch and autodialer, with storage for 200 telephone numbers. The talking S-meter lets the user check signal strength into the repeater. The controller also supports pocket pagers, linking to other repeaters, and a bulletin board. It has high-quality synthesized voice with ACC's large, custom speech vocabulary.



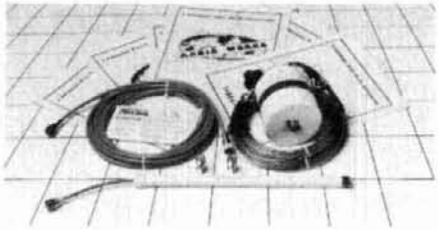
The '96 has built-in keypad and indicators, shielded DIN cables for easy hookup, and easily accessible pots and DIP switches in the back. The risk of lightning damage is minimized by a gas discharge tube across the phone line and transient suppressors on each I/O signal.

Contact Advanced Computer Controls, Inc., 2356 Walsh Avenue, Santa Clara, California 95051.

Circle #306 on Reader Service Card.

the Carolina Windom

The Radio Works has introduced the new Carolina Windom©, a high performance, 80-10 meter antenna system. While not a windom in the classic sense, its off-center feed system suggests the name. Fed with 50-ohm coax, it produces a low SWR across nearly all of the 75/80 meter band. Operation on 40-10 meters requires a transmatch.



The Carolina Windom comes with a special dedicated matching unit, vertical radiator section, high power line isolator, No. 14 stranded antenna wire, and glass-filled insulators. The package includes CoaxSeal® and an illustrated manual.

The price is $75, complete and ready to install. For more information, contact Jim Thompson, W4THU, Radio Works, Box 6159, Portsmouth, Virginia 23503. A catalog offering a wide selection of wire antennas, parts and accessories is available on request.

Circle #307 on Reader Service Card.

MAXFAX™ and WEFAX

Kantronics has added a weather facsimile command, WEFAX, with EPROM update 2.8. This update is available for the KAM, KPC-4, KPC-2, KPC-1, and the KPC-2400. In addition, Kantronics introduces two programs to work with the KAM or KPCs, MAXFAX-64/128 for Commodores and MAXFAX-PC for PCs and compatibles. If you use a PC, the CGA (color graphics adapter) is required.

With MAXFAX, you can store the pixel bytes from the KAM or KPC directly in RAM to the screen, or from RAM to diskette for transport or to your graphics printer. An Epson graphics format such as the EPSON LX-80 is assumed. Each MAXFAX copy comes on diskette and costs $19.95. You can order from Kantronics, Inc., 1202 E. 23rd Street, Lawrence, Kansas 66046.

Circle #308 on Reader Service Card.

new rf/high-voltage adapters

Nemal Electronics International offers a new line of rf adapters for quick and reliable interconnections between different connector series. These adapters facilitate rapid interconnection of incompatible cables and equipment while maintaining low loss and VSWR.

The NE964 adapts a type HN jack to an SMA plug; the NE962 adapts an SMA jack to a type HN plug; and the NE970 adapts a type N jack to an LC plug. The NE866 is a BNC series bulkhead feedthrough with both isolation and hermetic seal.

Constructed of brass and plated silver/nickel, these adapters have Teflon™ insulation and tolerate temperatures of −55C to +199C. Other specifications include an impedance of 50 ohms, frequency of 0-4 GHz, VSWR of 1.3, and voltage rating of 375/1500. The adapters offer electrical performance and construction to military specifications.

For details contact Nemal Electronics International Inc., 12240 NE 14th Ave., North Miami, Florida 33161.

Circle #309 on Reader Service Card.

AR-501 radio telegraph terminal

ACE Communications, Inc.'s model AR-501 is a triple-mode radio telegraph (CW) terminal for Amateur Radio operator, and short wave listeners.

The AR-501 performs as a CW decoder, CW trainer, and electronic keyer. Features include: automatic speed follow-up and threshold control, LED tuning indicator, 32-character LCD display, random code generator, and electronic keyer for both standard and iambic. Codes can be monitored in all three modes by internal speaker and printer through the parallel printer port.

It measures 4.5 x 6.25 x 2.25 inches and is powered by a 12 VDC source. The price of the AR-501 is $229.00 including ac power adapter and parts for hookup.

For more details contact ACE Communications, Inc., 22511 Aspan St., El Toro, California 92630-6321.

Circle #310 on Reader Service Card.

high-quality variable capacitors

Kilo-Tec announces the availability of the Nevada High-Power variable capacitors. They are capable of withstanding very high rf voltage up to 7.8 kV. These heavy-duty caps are suitable for high-power antenna matching units, power amplifiers, and transmitters.

There are two values: a 500 pF and a 250 pF. Approximate prices are $29 for the TC-250 and $40 for the TC-500. To order or receive a quote contact Kilo-Tec, PO Box 1001, Oak View, California 93022.

Circle #311 on Reader Service Card.

DX FORECASTER

Garth Stonehocker, KØRYW

sporadic E season — 1988

This is the second summer after the end of solar cycle 21. What kind of sprodic E conditions can we expect? In the period from May through September radiation from the nearly overhead sun generates high ion densities in the lower ionosphere that support short-skip propagation, including multiple short skips. The geomagnetic field clusters these ions into cloud-like patches known as sporadic-E (E_S). These patches form a thin layer of intense ionization in the E region about 60 miles up. A patch gives a strong, mirror-like signal reflection over skip distances of 600 to 1200 miles. Signals remain strong for about half an hour, up to a couple of hours after the onset of the first strong signal.

The frequency and magnitude of Sporadic E occurrences is a function of geographical location. The best locations for E_S openings this summer are toward the equator and on either side of the geomagnetic equator. It's especially good where the geomagnetic equator is furthest from the geographic equator. The Northern Hemisphere areas are: Southeast Asia (best) and the Mediterranean (next best) followed by South America in the Southern Hemisphere. These were shown graphically in this column last year on a contour map.

The highest frequency propagated by E_S tends to occur at local noon. The E_S patch is imbedded in the regular E layer and tends to track the E maximum ion density throughout the day, season, and sunspot cycle. During this summer expect about a 17 percent increase in the E layer as an E_S base for higher maximum usable frequencies (MUF) over a 1200 mile hop. This increase gives the base MUFs of 47 to 53 MHz this year, so six meter openings should be more prevalent. Two meter openings may still be rare, especially this month; perhaps August will provide some. The highest probability of occurrence is near sunrise and again around sunset. These two E_S characteristics affect short-skip openings differently. Openings on the higher-frequency bands occur around local noontime; the lower bands tend to have openings near sunrise and sunset. This occurrence characteristic is nearly constant over the sunspot cycle so there should be the same number of low to midlatitude E_S openings in the next few years.

last-minute forecast

Expect the higher frequency DX bands to be very good during the first two weeks of June because of solar flux peaks and longer daytime hours. Both factors contribute to elevated MUFs during the evening at midlatitude locations. No single hop transequatorial openings are expected but look for good sporadic E openings around noon toward the end of the month. Good nighttime DX conditions on the lower bands are expected during the last two weeks of the month, but they will be noticeably shorter in duration and noisier as northern tropical thunderstorm noise propagates toward us. Geomagnetic disturbances are anticipated from solar flares around the 5th, more probable on the 13th, and from coronal thinning on the 18th through 24th of the month. MUFs should decrease about 15 percent on east-west propagation paths on most days and probably 20 percent on those paths during disturbed conditions on the 13th. Signals should be 10 to 15 dB lower level and QSB will be noticed. Paths near the equator can expect 10 percent higher MUFs.

The moon will be full on June 29th and at perigee (its closest approach) on June 5th. Summer solstice is on the 21st at 0357 UTC. The Aquarid meteor shower starts about the 8th, peaks around the 28th, and lasts until about August 7th. The maximum radio-echo rate will be 34 per hour.

band-by-band summary

Six meters will provide occasional openings to South Africa and South America around noontime via short-skip E_S propagation.

There will be long-skip conditions on ten meters in the afternoon during the peak times of the 27-day solar cycle. Otherwise, look to sporadic-E short-skip and multihop openings around

JUNE

WESTERN USA

	2300	2200	2100	2000	1900	1800	1700	1600	1500	1400	1300	1200	1100	1000	0900	0800	0700	0600	0500	0400	0300	0200	0100	0000	GMT
JUNE	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	PDT
ASIA FAR EAST	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	20*	20	20	20	20	N
EUROPE	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	NE
S. AFRICA	15	15	12	10	10	10	10	10	12	12	15	15	15	20	20	20	20	20	20	20	20	20	15	15	E
S. AMERICA	12	12	12	15	15	15	20	20	20	20	20	20	20	20	20	15	15	12	12	10	10	10	10	12	SE
ANTARCTICA	15	15	20*	20	20	20	20	20	30	30	30	30	30	30	30	30	30	30	30	30	30	20	20	20*	S
NEW ZEALAND	10	12	12	12	15	15	20	20	20	20	20	20	20	15	15	15	12	12	12	10	10	10	10	10	SW
OCEANIA AUSTRALIA	10	12	12	15	15	15	20	20	20	20	20	20	20	15	15	15	12	12	10	10	10	10	10	10	W
JAPAN	15	20	20	20	20	20	20	20	30	30	30	20	20	20	20	20	15	15	15	15	15	15	15	15	NW

MID USA

	2300	2200	2100	2000	1900	1800	1700	1600	1500	1400	1300	1200	1100	1000	0900	0800	0700	0600	0500	0400	0300	0200	0100	0000	GMT
	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	MDT
ASIA FAR EAST	20	20	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	N
EUROPE	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	NE
S. AFRICA	15	12	12	10	10	10	10	10	12	12	15	15	15	20	20	20	20	20	20	20	20	20*	15	15	E
S. AMERICA	15	15	12	12	10	10	10	10	12	12	15	15	15	20	20	20	20	20	20	20	20	20*	15	15	SE
ANTARCTICA	20	20	20	20	20	20	20	20	30	30	30	30	30	30	30	30	30	30	30	30	30	20	20	20	S
NEW ZEALAND	10	10	12	12	12	15	15	20	20	20	20	20	20	20	20	15	15	12	12	10	10	10	10	10	SW
OCEANIA AUSTRALIA	10	10	12	12	12	15	20	20	20	20	20	20	20	20	15	15	12	12	12	10	10	10	10	10	W
JAPAN	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	20	20	NW
	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	CDT

EASTERN USA

	2300	2200	2100	2000	1900	1800	1700	1600	1500	1400	1300	1200	1100	1000	0900	0800	0700	0600	0500	0400	0300	0200	0100	0000	GMT
	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	EDT
ASIA FAR EAST	20	20	15	15	15	15	15	15	20*	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	N
EUROPE	20	20	20	20	20	20	20	20	20	20	20	30	30	30	30	30	30	30	20	20	20	20	20	20	NE
S. AFRICA	15	15	12	10	10	10	10	12	12	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	E
CARIBBEAN S. AMERICA	15	15	12	12	10	10	12	12	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	SE
ANTARCTICA	20	20	20	20	20	20	20	20	30	30	30	30	30	30	30	30	30	30	30	30	30	20	20	20	S
NEW ZEALAND	10	10	12	12	12	20	20	20	20	20	20	20	20	20	20	20	15	15	12	10	10	10	10	10	SW
OCEANIA AUSTRALIA	10	12	12	12	15	20	20	20	20	20	20	20	20	20	20	15	12	12	10	10	10	10	10	10	W
JAPAN	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20	20*	15	15	15	15	15	15	20*	NW

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides MUF during "normal" hours.

*Look at next **higher** band for possible openings.

local noon for DX on this band. (Evening transequatorial openings usually don't occur in the summertime.)

Twelve and fifteen meters, almost always open to some southern part of the world, will be the main daytime DX bands. Operate on 12 first, then move down to 15. DX is considered 5000 to 7000 miles on these bands. There may be some long, one-hop transequatorial propagation paths occurring early in the month.

Twenty, thirty, and forty meters will support DX propagation from most areas of the world during the daytime and into the evening hours most days. Forty meters joins this daytime DX group because of lower signal absorption, and therefore lower LUF (lowest usable frequency) during this last sunspot minimum year. DX on these bands may be either long-skip to 2500 miles or short-skip E_s to 1250 miles per hop. There are many good hours of DXing ahead as the days get longer.

Thirty, forty, eighty, and one-sixty are all good for nighttime DX. Although the background thunderstorm noise is becoming noticeable, these bands are still quiet enough to provide good DX working conditions. Sporadic-E propagation may be a contributing factor toward enhanced conditions at local sunset and will occur more often during the next three months.

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MACINTOSH SATELLITE Tracking and Propagation Software. MacTrak™ displays graphic maps (rectangular, polar, great circle), views from space, schedules, windows. Also tracks Sun and Moon. Compatible with KLM/MIRAGE rotor interface. Shows gray line, sunrise/sunset, bearing/distrnac,e mUF vs time, areas of world "open". SASE for info or $49.95 from R. Stegemeyer, PO Box 1590, Port Orchard, WA 98366.

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SELL: AEA PK64A/HFM packet $175. ICOM IC 271A 25W, 2m Xcvr $525. Hallicrafters SX115 Receiver $250 Spectrum Communications 450/2m repeater circuit boards TX/RX $50 each and RF power amplifiers $35 each. VHF Engineering ID $25 and COR $10. Radio Shack Color 3 Computer with RS-232 interface $150. ICOM 251A 2m, 10W multimode Xcvr $300. Floyd Chowning, K5LA, 5637 Prince Edward Avenue, El Paso, Texas 79924. (915) 751 6204.

HAMLOG COMPUTER PROGRAMS. 17 modules auto logs, sorts 7-band WAS/DXCC. Full features. Apple $19.95, IBM or CP/M $24.95. KA1AWH, POB 2015, Peabody MA 01960.

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RECONDITIONED TEST EQUIPMENT $1.25 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

COMING EVENTS

Activities — "Places to go . . ."

SPECIAL REQUEST TO ALL AMATEUR RADIO PUBLICITY COORDINATORS: PLEASE INDICATE IN YOUR ANNOUNCEMENTS WHETHER OR NOT YOUR HAMFEST LOCATION, CLASSES, EXAMS, MEETINGS, FLEA MARKETS, ETC. ARE WHEELCHAIR ACCESSIBLE. THIS INFORMATION WOULD BE GREATLY APPRECIATED BY OUR BROTHER/SISTER HAMS WITH LIMITED PHYSICAL ABILITY.

WASHINGTON: June 4 and 5. The Apple City Radio Club's Hamfest, Rocky Reach Dam, 7 miles north of Wenatchee, Hwy 97. Saturday license exams. Free camping and RV spaces. Advance registration $4.00 to Bob Lathrop, K7EVL, 919 N. Woodward, Wenatchee, WA 98801.

MICHIGAN: June 11. The 14th annual Hamfest sponsored by the Central Michigan Amateur Repeater Association (CMARA), Midland Community Center, Midland. 8 AM to 1 PM. Donation $3. New/used Amateur electronics and computer equipment. License exams. Dealers welcome. Talk in on 147.60/000, Midland. For more information please SASE to CMARA Hamfest, PO Box 67, Midland, MI 48640. (517) 631-9228 evenings and weekends.

NORTH CAROLINA: June 11. Winston-Salem Hamfest & Computer, Electronics Fair '88. Sponsored by the Forsyth ARC. Dixie Classic Fairgrounds. FCC exams, walk-ins welcome. Free parking, convenient lodging, indoor dealers, large flea market/tailgating area. For dealer information write Forsyth ARC, Hamfest Committee, PO Box 11361, Winston-Salem, NC 27116.

ILLINOIS: June 12. The 31st Annual Hamfest sponsored by the Six Meter Club of Chicago, Santa Fe Park, 91st and Wolf Road, Willow Springs. Advance registration $3.00. At the gate $4.00. Large Swappers Row, Pavillion Displays, AFMARS meeting, plenty of parking, picnic grounds. No overnight camping. Advance tickets from Mike Corbett, K9ENZ, 606 South Fenton Avenue, romeoville, IL 60441 or any Club member. Talk in on K9ONA 146.52 or K9ONA/R 37-97.

QUEENS, NEW YORK: June 12. The Hall of Science ARC Hamfest, NY Hall of Science Parking Lot, Flushing Meadow Park, 4701-111 Street, Queens. 9 AM to 3 PM. Amateur Radio exhibit station. Tune up clinic, films. Buyers $3.00. Sellers $5.00 per space. Talk in on 144.300 simplex link 223.600 rep and 44.225 rep. For further information call Steve Greenbaum, WB2KDG (718) 898-5599 or Arnie Schiffman, WB2YXB (718) 343-0172 evenings.

MASSACHUSETTS: June 12. SEMARA Hamfest sponsored by the SE Massachusetts ARA, South Dartmouth. 9 AM to 5 PM. Admission Free. Dealers $8.00 advance. Door $10. VE exams appointment only. "Nepra" Packet Workshop. Tailgating. Talk in on 147.000 and .6 for Hamfest and 145.4900 .6 for back up. For VE exams and Hamfest info SASE to Pete Kodis, N1EXA, PO Box 9187, North Dartmouth, MA 02747.

KENTUCKY: June 12. "HAM-O-RAMA 88" sponsored by the Northern Kentuck ARC, Erlanger KY Lions Park. Admission $4.00 advance; $5.00 at the door. Children under 13 free. ARRL, Packet and Emergency Forums. Indoor exhibit area; $15.00 per space. Large outdoor flea market, $4.00 per space. Setup 6 AM. General admission 8 AM. Food and refreshments available. Talk in 147.855/.255 and 147.975/.375. For more information or advance registration contact WA4BRM, c/o NKARC, PO Box 281, Florence, KY 41042. (606) 371-8545.

NEW JERSEY: June 18. The Raritan Valley Radio Club's 17th annual Hamfest, Columbia Park, Dunellen. Gates open 8 AM. Lookers $4.00 donation. Spouse and kids free. Sellers $6.00/one space. $12.00/multiples. Bring your own tables. Talk in on club repeater W2QW/R 146.025/.625 and 146.52 simplex. For information call Dave, KA2TSM (201) 763-4849 or John, WA2C (201) 968-5070 or any club member.

MICHIGAN: June 18. The Independent Repeater Association is sponsoring its annual Hamfest, National Guard Armory, 44th Street, 1/2 mile west of US 131, Grand Rapids. 8 AM to 4 PM. Dealer setup 6 AM. Free reserved tables for dealers and sellers. Talk in on 147.165/147.765. For table reservations: Independent Repeater Association, 562 92nd Street SE, Byron Center, MI 49315. (616) 455-3915.

CALIFORNIA: June 19. The Satellite ARC's annual Father's Day Swapfest, Union Oil Company New Love Picnic Grounds south of Santa Maria on US 101. Free Admission. Starts 9 AM. Bar-B-Que served at 1 PM. Tickets: Adults $6.50, children 6 to 12 $3.50, under 6 free. Swap tables available. Talk in in 145.14 repeater. Simplex 146.52. For tickets and information Hank Korczak, W6PME, Santa Maria Swapfest, 917 Anthony Way, Lompoc, CA 93436 (805) 736-1761.

MARYLAND: June 19. The Frederick ARC will hold its 11th annual Hamfest, Frederick County Fairgrounds. 8 AM to 4 PM. Admission $3.0. Tailgaters $2.00 extra. Exhibitors tables $10/first, $5/each extra. Setup 8 PM June 18. For information write Dave Durkovic, N3BKD, 7128 Limestone Lane, Middletown, MD 21769.

PENNSYLVANIA: July 4. The Harrisburg Radio Amateur Club is sponsoring a Hamfest, Bressler Picnic Grounds. 8 AM. Vendors 7 AM. Admission $3.00 at gate. Tailgating $1.00 per site. Pavillion tables $5.00 each. Campground, motels and restaurants nearby. Contact Dave Dormer, KC3MG, 131 Livingston St, Swatara, PA 17113. Phone (717) 939-4957.

WISCONSIN: July 9. The South Milwaukee ARC will hold its annual SWAPFEST, American Legion Post 434, 9327 South Shepard Avenue, Oak Creek. 7 AM to 3 PM. Admission $3.00. Free overnite camping available. Picnic area. The Badger Exams group will conduct exams during the day. Talk in on 146.580 simplex. For more details including a map write The South Milwaukee ARC, PO Box 102, South Milwaukee, WI 53172-0102.

INDIANA: July 9 and 10. The 18th annual State ARRL Convention and Hamfest, Marion County Fairgrounds, Indianapolis. Gates open 6 AM both days. Gate fee $5.00. Under 12 free. Dealers, electronic flea market, homebrew, tech and ARRL forums. Non-ham activities. Free hookup and camping, food available on grounds. Neary motels and restaurants. For information: Flea Market Space (317) 356-4451. Commercial (317-745-6389.

PENNSYLVANIA: July 10. North Hills ARC's 3rd annual Hamfest, Northland Public Library, 300 Cumberland Road, Pittsburgh. 8 AM to 4 PM. Free admission. Free dealer and flea market space. VEC exams. ARRL table. Refreshments. HANDICAP AND WHEELCHAIR ACCESSIBLE. For Hamfest info: Bob Ferrey, Jr, N3DOK, 9821 Presidential Drive, Allison Park, PA 15101. (412) 367-2393. VEC info: John Rosenwald, NM3P, 400 Stevens Drive, Pittsburgh, PA 15237. (412) 931-2631.

WEST VIRGINIA: July 17. The 10th annual TSRAC Wheeling Hamfest/Computer Fair, Wheeling Park, 9 AM to 4 PM. WV's Largest. Dealers welcome 30,000 square feet under roof; 5 acres flea market. Family activities at Park. Admission $3.00 in advance. $4.00 at door. To reserve space contact: Sandi Williams, KB8AAV, 9 East High St, Flushing, OH 43977; for tickets, TSRAC, Box 240, RD 1, Adena, OH 43901 (614) 546-3930.

OPERATING EVENTS

"Things to do . . ."

June 4 and 5: The Wireless Institute of Northern Ohio (WINO) will be on the air with a special event station to commemorate Ohio Wine Month. Saturday evening 7 to 11 PM EDST, and Sunday 11 AM to 3 PM. The station will be located at a Madison, Ohio winery and will use callsign KO8O. For a special 8x11 certificate send legal sized SASE to KO8O, WINO Weekend, 10418 Briar Hill, Kirtland, Ohio 44094.

June 19 to 25: E.A.R.S. will operate N6WB in conjunction with the 100th Anniversary of the City of Escondido, CA. For a large certificate send QSL and large SASE to Glenn Bodeker, N6WB, 127 Walnut Hills Drive, San Marcos, CA 92069.

June 20 to July 4. The 6th annual Great American Race, Disneyland to Boston. KX6B will operate mobile as part of the support team of Car No 73, to commemorate the 80th year since the original New York to Paris Great Race. Operation daily 1500 UTC to 2400 UTC. Lower 25 kHz of 40, 20 and 15m General phone bands. 10m Novice phone. Mobile Packet 145.01 and 2m FM. For QSL SASE to Dick Raley, KX6B, 2610 Camloop Drive, San Jose, CA 95130.

July 4 and 5: The Valley ARA, Staunton, Virginia, will operate N4ICT in conjunction with the Statler Brothers Happy Birthday USA Celebration. Freqs: Phone 3.855, 7.280 14.250 and 28.375. For the Statler Brothers certificate send QSL, contact number and 9x12 SASE to Valley ARA, PO Box 666, Staunton, VA 24401.

THE FOUNDATION FOR AMATEUR RADIO, INC plans to award twenty-eight scholarships for academic year 1988-89 to assist licensed Radio Amateurs who plan to pursue a full time course of study beyond high school and are enrolled or have been accepted by an accredited university, college or technical school. For further information write FAR Scholarships, 6903 Rhode Island avenue, College Park, MD 20740 prior to May 31, 1988.

HAM EXAMS: The MIT UHF Repeater Association and the MIT Radio Society offer monthly Ham Exams. All classes Novice to Extra. Wednesday JUNE 22. 7 PM, MIT Room 1-150, 77 Mass Ave, Cambridge, MA. Reservations requested 2 days in advance. Contact Ron Hoffmann at (617) 646-1641. Exam fee $4.50. Bring a copy of your current license (if any), two forms of picture ID, and a completed form 610 available from the FCC in Quincy, MA (617) 770-4023.

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ELMER's NOTEBOOK

Tom McMullen, W1SL

"Q" signals

There are times when we are reminded that not everyone is familiar or experienced in the language of Amateur Radio. I was following my usual practice of listening to the activity on a local repeater while on my way to work one morning, and encountered a lively discussion on the meaning of a "Q" signal.

The signal in question was "QRU," and each Amateur knew part of the answer and thought that the other was incorrect. I resisted the temptation to break in and enlighten them about the "true meaning and proper use," but instead waited to see what developed. Sure enough, the next morning, the pair got together again; one had looked it up and was now fully informed. He surprised the other Amateur by stating that they were both on the right track, but needed the whole story. As with most Q signals, QRU can be either a question or a statement. When followed by a question mark, it (naturally) becomes a question. Without the question mark, it is a statement or an answer to a question. The discussion and follow-up not only educated the two Amateurs directly involved, but was also helpful to the many ears tuned to that repeater on those two days. Further, the incident triggered a thought that I'm putting to use here — how many other Q signals are unknown or misunderstood by a large number of Amateurs?

why Q signals?

Everyone uses Q signals. Old-timers cringe when hearing Q signals used in voice communications. Their theory is that such signals were invented for CW use, and if you are talking, you should say the phrase instead of the abbreviation.

There was a time when, deeply involved in traffic handling on several CW nets around the country, I agreed with that philosophy. However, after several years of exposure to the voice (and digital) world, I can see the merits of using Q signals wherever they apply, on any mode of communications.

Q signals, and their early companions, "Z" signals, were developed as short-cuts in message-handling procedures in marine and commercial radio circuits. It certainly was easier and quicker for an operator to send "QRU?" instead of "Do you have any messages for me?" The answer, equally shortened, would be either "QRU" (I have nothing for you), or "QTC" (I have messages for you). Before you old-time traffic handlers jump on me, yes, I've tweaked the phrase a bit. QTC really stands for "I have ... telegrams for you," but Amateurs are not in the business of sending telegrams. Anyway, the short Q signal reduced the amount of key-pounding, and to a busy commercial operator, this was a blessing. Amateurs, too, realized the advantage in both time and clarity in using abbreviations and operating signals, and adapted many of them to fit their operations. The "Z" signals served the same purpose in many commercial circuits, but for some reason never caught on with the Amateur fraternity — perhaps because ARRL (American Radio Relay League) publications listed and explained the use of Q signals. Also, it has been rumored that Z signals were proprietary to some network or service, but I've not been able to find a reference that proves this.

voice and digital useage

Everyone uses Q signals on voice operation from time to time. The old

Table 1. Common Amateur Q signals

QRM	Is there interference on the frequency?	There is interference on the frequency
QRN	Is atmospheric noise (static) bothering you?	Atmospheric noise (static) is bothering me.
QRP	Shall I reduce power? (Seldom used by Amateurs as a question.)	Reduce power (Most often used as a statement, as in "I am running QRP here" meaning the power is only a few watts.)
QRS	Shall I send slower?	Send slower.
QRT	Shall I stop sending?	Stop sending. (Usually used to mean the station is shutting down for the moment, as in "I'm going QRT for now.")
QRU	Do you have anything for me?	I have nothing for you.
QRV	Are you ready?	I am ready.
QRX	Shall I wait?	Wait (most often used as in "QRX 5 minutes.")
QRZ	Who is calling? (This is not a substitue for "CQ".)	....is calling you. (Amateurs seldom use QRZ as a reply.)
QSB	Does my signal strength vary?	Your signal strength varies.
QSL	Do you acknowledge?	I acknowledge.
QSO	Are you in contact with ....? (Amateurs seldom use QSO as a query.)	I am in contact with, or, I have made contact with ..... (More often used in referring to a contact between two Amateurs, as in "Thanks for the QSO, and 73 to you.")
QSY	Shall we change frequency	Let's change frequency ....
QTH	What is your location?	My location is ....

standard "QSL?" is used to mean several things: "Do you copy?", "Did you copy?", "Do you understand?", and so forth. The answering statement, "QSL" applies to all these questions and more.

When conditions are good, and the signals are "arm-chair copy" between the two stations, there's really no justification for using a voice Q signal, but habits don't get turned on or off according to band conditions. When conditions are poor, or there is abundant interference (there it is again — the Q signal QRM applies), certainly the letter Q sets the listener up to expect two more letters that are pertinent to the situation, and it might be easier to understand "QSY up 3" than "Let's move up 3 kilohertz".

In digital communications, the need is not so much for overcoming interference or weak-signal conditions — packet and AMTOR systems handle that pretty well — but rather a way to reduce the keystrokes at the sending station. Not all packet and RTTY enthusiasts are expert typists, and a few 3-stroke Q signals that can take the place of a whole line of text are a blessing to both sender and receiver. (How often I've stared at a blank screen wondering if something was not working right, only to find the operator at the other end was "one-finger typing" the message.)

In summary, Q signals are both useful and permissible in any mode today. It will help Novices and higher-class licensees to feel more at home on the air if they know what Q signals to use and how to use them. **Table 1** lists the most common signals in both their question and answer form. This is by no means a complete list — some, like "QTE?" (What is my true bearing in relation to you?") would probably be hard to understand and elicit a "HUH?" (which, fortunately, needs no Q signal).

I have modified the original meaning of many of these signals a bit, to make them more compatible with current Amateur Radio usage. The original Q signals were developed for commercial and aircraft use, and the language was either more stilted or directly applicable to a specific situation. As they are wont to do, Amateurs have softened the language and slanted the meaning to fit their needs, which **table 1** reflects.

Amateur traffic nets, both CW and voice, have their own set of Q signals that help to speed up message handling and network management. Many are adaptations of more common signals, with the middle letter replaced by an "N," as in QNU, which is borrowed from QRU, meaning "I have no traffic for the net." Another net signal is QNX, meaning "You are excused from the net." A few minutes spent listening to some of the busier traffic nets on 80-meter CW, 75-meter phone, and a few 2-meter repeaters is a lesson in management and a discipline that gets things done efficiently. When you read the monthly message totals as reported in *QST*, you can see why.

There's another signal — QST. It does not have a question as part of its definition. QST is an alerting call to all Amateurs, indicating that some important information is to follow. It can be used by anyone, and is often heard at the beginning of network announcements and 2-meter repeater emergency-practice sessions. You're undoubtedly familiar with its use before code practice and bulletin transmissions from W1AW, the ARRL Maxim Memorial Station in Newington, Connecticut, and on the cover of their magazine, *QST*, which is the official journal of the American Radio Relay League.

Q signals are a vital and interesting part of Amateur language, useful in conveying information quickly and showing that you are "with it" on the bands. They fit all modes of communication (yes, even Amateur Television — a snowy picture of a card that says QRX 5 in big letters will get its message across), and when both the sender and the receiver know the meaning of "QRM, QSY down 3," things work a lot smoother!

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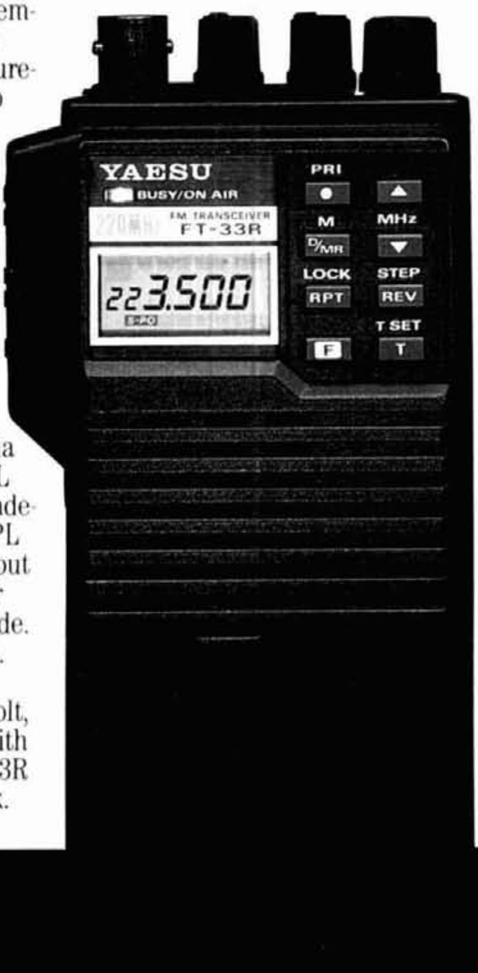

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